RADIO AND ELECTRONIC
LABORATORY HANDBOOK

RADIO AND ELECTRONIC LABORATORY HANDBOOK

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(Key subjects are shown in bold type)

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Preface

During the twenty-three years since the first edition of this book the situation has changed drastically. Whereas in 1938 there was so little choice of ready-made equipment that the experimenter had to construct much of what he wanted, now the variety offered by the instrument trade is embarrassingly large. Both the instruments and the ways in which they are used have become much more sophisticated. Experimenting is a good deal less light-hearted than it used to be, and far more people are doing it—mostly as professionals rather than as amateurs.

So this seventh edition is hardly recognizable, either in size or content, as a descendant of the first. Even the title is slightly different this time, in recognition of the fact that the techniques are now applied far beyond the field of radio. But all the editions have had the same basic aim—to provide the experimenter with guidance on means and methods, without assuming unlimited funds and space at his disposal, or substantial qualifications and experience. Certain chapters—notably 3, 8 and 13—deal with the underlying principles of experimenting and the approach to it, seldom discussed or considered, but vitally important. It is hoped, too, that the book will not disappoint those readers who have been good enough to say they liked the original style—aimed to be a shade less solemn than the more advanced textbooks. Students—and, in these days, sixth-formers—and sometimes even their teachers, seem to appreciate this.

Possessors of the sixth edition may need to be convinced that it is going to be worth while changing over to this one. It should only need to be mentioned that in the interval transistors have come into general use, and that there has been a rapid trend towards pulse waveforms, resulting in radical redesign of oscilloscopes and other indicating instruments. So the chapters on instruments and many sections of those on measurements have been largely rewritten.

A still more determined effort has had to be made to keep the book within reasonable bounds, by deletion and by selecting from the vast range of instruments and methods those likely to be most generally useful and best value for money, and presenting the information as concisely as possible consistent with the aim of helping those with limited experience and learning. In many instances, references to the literature for details have been regrettably necessary. Microwaves have been excluded; they now have a large literature of their own.

To avoid having to repeat details of apparatus in every method employing it, the chapters on apparatus come first. General principles and reference material are stored at the end in Chapter 14.

Any necessary link-up is facilitated by cross-references: all sections, figures and tables are numbered, the number of the chapter being given first; e.g. Sec. 14.23 is the 23rd section in Chapter 14. But the index and table of contents should not be forgotten; they are there to be used.

The basic units are those of the rationalized m.k.s. system (Sec. 14.1). Whenever other units are used they are specified in square brackets; e.g.,

$$C = \frac{25{,}330}{f_r^2 L} \qquad [\mathrm{pF};\ \mu\mathrm{H};\ \mathrm{Mc/s}]$$

Symbols and abbreviations are British Standard, set out in detail in Secs. 14.1 and 2. Full use is made of the nanofarad (nF), equal to 1,000 pF or 0·001 μF. A large proportion of the capacitances used in radio are of the order of 1 nF, and avoidance of the many confusing decimal places would alone justify it. But by enabling nearly all values to be given in small whole numbers it removes the last excuse for the time-wasting practice of specifying component values in a separate list instead of on the circuit diagram itself. The only necessary indications of unit are the multipliers (p, n, μ, m, k, or M), and by using these so as to avoid unnecessary noughts, and leaving the circuit symbol to denote the main unit (F, H, or Ω), the values can be shown clearly on the most complicated circuit diagram. Quantities are denoted by italics; e.g., *R* for resistance, in distinction from R for a resistor.

Another usage, which may not yet be familiar to everyone, is the term " z.f.", meaning " zero frequency ". Attempts to do without it result in absurdities such as " d.c. current " and " at d.c."

References to literature are admittedly not in standard form as laid down in B.S. 1629: 1950. This is because it is believed that the first concern of most readers is not the author but the subject, and secondly that the date of a paper is more helpful than the volume number. Any injustice to authors in this arrangement has, it is hoped, been removed by placing all their names in the index. The titles of books and periodicals are given in italics; most of the latter are abbreviated as shown opposite. Publishers of books are indicated in brackets, and their full names and addresses in Britain are also shown opposite.

Thanks are due to the firms (named in the captions) which have contributed photographs illustrating technical features of equipment, and to those that have supplied information on their products. The encouragement to write the first edition given by Mr. C. R. Cosens of Cambridge University, as well as his suggestions since, are gratefully acknowledged. Considerable help has been given in the preparation of this seventh edition by Mr. G. W. Short.

Finally, the author will much appreciate reasoned criticisms or the pointing out of any errors that may have occurred.

18, *Bromley Common*, M.G.S.
Bromley, *Kent*.
October, 1961.

Full Titles of Journals

B.C. & E.	British Communications and Electronics.
E. & R.E.	Electronic & Radio Engineer.
E.E.	Electronic Engineering.
E.T.	Electronic Technology.
G.R.E.	General Radio Experimenter.
J. Brit. I.R.E.	Journal of the British Institution of Radio Engineers.
J.I.E.E.	Journal of the Institution of Electrical Engineers.
J. Sci. Inst.	Journal of Scientific Instruments.
M.I.	Marconi Instrumentation.
M.T.C.	Mullard Technical Communications.
Proc. I.E.E.	Proceedings of the Institution of Electrical Engineers.
Proc. I.R.E.	Proceedings of the Institute of Radio Engineers (New York).
Proc. Wireless Section I.E.E.	Proceedings of the Wireless Section of the Institution of Electrical Engineers.
S.R. & R.	Sound Recording and Reproduction (Journal of the British Sound Recording Association).
W.E.	Wireless Engineer.
W.W.	Wireless World.

Publishers

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Allen & Unwin	George Allen & Unwin, Ltd., 40, Museum Street, W.C.1.
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Constable	Constable & Co., Ltd., 12, Orange Street, W.C.2.
Griffin	Charles Griffin & Co., Ltd., 42, Drury Lane, W.C.2.
H.M.S.O.	Her Majesty's Stationery Office, York House, Kingsway, W.C.2.
Iliffe	Iliffe Books, Ltd., Dorset House, Stamford Street, S.E.1.
Longmans	Longmans, Green & Co., Ltd., 6, Clifford Street, W.1.

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McGraw-Hill	McGraw-Hill Publishing Co., Ltd., 95, Farringdon Street, E.C.4.
Macmillan	Macmillan & Co., Ltd., St. Martin's Street, W.C.2.
Methuen	Methuen & Co., Ltd., 36, Essex Street, W.C.2.
O.U.P.	Oxford University Press, Amen House, Warwick Square, E.C.4.
Pitman	Sir Isaac Pitman & Sons, Ltd., 39, Parker Street, W.C.2.

CHAPTER 1

The End and the Means

1.1. PURPOSE OF A LABORATORY

The way a laboratory is equipped and carried on depends very largely on the purpose it is meant to serve. Within the general field there are a number of quite distinct purposes, which perhaps can be conveniently identified by the words Demonstration, Research, Design, Test and Maintenance.

Demonstration.—This is to repeat the results of others and so gain first-hand knowledge. While this suggests the technical school with its set experiments, it is a still more valuable object for private work. An ounce of practice is worth a ton of theory, we are told. While this exact numerical ratio does not appear to have been rigorously proved, there is no doubt that personally confirming experiments does give one a mastery of the subject that can never be derived from books and lectures only. Moreover it may lead to

Research, or finding out more about how things work and so possibly discovering something new. Some of the most valuable discoveries have been made as a result of *not* confirming the findings of others. " Take nothing for granted " is too strict a rule to observe literally, but the spirit of it is sound. Many lines of research accidentally grow out of and perhaps utterly transcend in importance the original work. Others, again, are deliberately planned and (sometimes!) carried through. Turning from these heights that are so fascinating to contemplate and often so tedious to climb, we find

Design.—Work under this heading really falls into two divisions: obtaining data for design; and checking the designs when they have materialized. This is the main purpose of most of the laboratories in manufacturing concerns, and on a smaller scale the same object is pursued by keen amateurs.

Test.—The private experimenter ought to make a practice of testing all the parts he obtains for his work. It is better to start doing this voluntarily than to be driven to it, sad but belatedly wise. In large organizations this work falls to the Test Room, but there are usually special tests that cannot or ought not to be done there. Tracing obscure faults in apparatus may be included in this division. Lastly comes

Maintenance.—This is probably common to all laboratories, from least to greatest; for there is always plenty of work to be done in keeping the instruments and other equipment in perfect order, making and checking calibrations—valve voltmeters, for example, ought to be checked quite often—and setting up new apparatus. Nothing is more exasperating, when carrying out an experiment, than continually having to digress in order to locate and cure faults in the apparatus, or (worse still) failing to notice such faults until one has taken a long series of readings, all of which must be condemned as unreliable.

1.2. THE QUESTION OF EQUIPMENT

The foregoing list, drawn up to help clarify ideas, is intended mainly for the experimenter who is setting up his own laboratory: the paid worker is generally not in a position to choose. When equipping a laboratory, one may save money by first considering rather carefully what one wants to do. The most lavishly-equipped plant without inspiration may be less truly a laboratory than the corner of a living-room with a few simple home-made instruments. At the same time, one cannot set out to experiment on long-wave directional aerial arrays in a small flat. But it is possible to make what might prove most valuable observations with little more than an ordinary receiver. The "Luxembourg Effect" (or ionospheric intermodulation, as it was called when it was at length admitted to serious science) was not discovered by means of elaborate and costly apparatus, nor was it predicted as a result of intricate mathematical calculations; the equipment was a receiver and slightly more than average observation. Yet here was a discovery that gave the scientists a valuable new line of approach to the problems of the ionosphere.

More recently, some strange effects in the propagation of television and other v.h.f. signals have been noticed. For investigating such things there is still room for work that takes the form of systematic observations, requiring little apparatus other than a good receiver and perhaps an output meter. But the emphasis should be on the word "systematic". The very short waves are especially suitable for study by the private investigator, for a number of reasons, among which may be mentioned the practicability of testing and comparing various types of resonant aerials without the need for several acres of land.

So far nothing very suggestive of laboratory apparatus has been mentioned. But even when the chosen experimental work can be carried on with only a receiver it is at least desirable to have something of the nature of a laboratory, if only for ensuring the very highest degree of reliability in that receiver. Work is very restricted without some means of producing an artificial signal when required, and measuring frequencies and component values. Meters for checking circuits and taking experimental readings, oscillators of various types, and controllable power supplies are constantly needed whenever more general experimental work is undertaken. And so one goes on. The question is how to lay out limited funds to the best advantage.

1.3. SELECTING FOR MINIMUM COST

The factor of cost will be taken into consideration when equipment is described in detail in subsequent chapters; but in the meanwhile it is possible to make some general observations.

Assuming that experimental work is not a fleeting passion, there are certain instruments that will be practically essential, both now and in the future, of a kind that does not quickly become obsolete. For these it is false economy to go in for the very cheapest that will just do at the time. What are loosely but very conveniently termed "meters" certainly come into this category. So do standards of resistance, capacitance, and so forth. Good workmanship now more than ever costs money, and in the ordinary way a high-grade instrument by a famous firm always commands a good price. In the *ordinary* way; but one does sometimes pick up exceptional bargains second-hand. And of course releases of Government surplus stocks are a great attraction to the bargain hunter. But in bargain hunting one must never forget the legal maxim "*caveat emptor*"—"Let the buyer beware!" As bargains do not generally carry any guarantee, such buying calls for discrimination, and it is advisable to be able to check them against standards. Unless one can get something much better this way than could be afforded otherwise, it is perhaps wiser to be on the safe side and buy a new instrument. Sometimes, however, the original makers, or a reputable agent, can supply used instruments that have been reconditioned and recalibrated and are practically equivalent to new, except that perhaps they are not of the latest pattern. But of course one cannot in this way expect to save a very large proportion of the cost of a new instrument.

1.4. TRUE ECONOMY

Decide, then, what apparatus is likely to be of permanent value, and get something of higher quality than appears to be justified, even if it means postponing other purchases. In doing so do not overlook the possibility of unit instruments, consisting of a nucleus to which accessories may be added as required.

Versatility is an important feature to be considered. Careful choice may enable a few instruments to take the place of many. But do not let catalogue descriptions rush you into buying a jack of all trades and master of none. And remember, too, that one instrument, however versatile, cannot be in more than one place in the circuit at a time.

With regard to more specialized types of equipment, of which there is now such a vast and rapidly increasing variety—waveform generators, distortion meters, transistor testers, and so on—more caution is necessary. They are constantly changing in design, are generally expensive, and tend to become obsolete fairly rapidly. This is not to say that a good signal generator, for example, is a bad investment; it is just that one needs to exercise foresight. People who a few years ago bought expensive television signal generators that stopped short of Band III find them very restricted to-day. Transistor testers become obsolete almost as soon as designed. It is bad policy to spend a lot on a splendidly finished and symmetrical instrument if the development of technique makes it necessary after a short time to disfigure it with some makeshift alteration.

1.5. MAKING ONE'S OWN APPARATUS

It is hereabouts that we enter the field of things that it may be advisable to construct rather than buy. One should not run away with the idea that it is necessarily much cheaper to do so. When the time spent on experimenting has been taken into account it may be quite an expensive way. But a private worker does not often rate his time very highly, and he learns all the time; and a technical department is often able to present the cost of construction to the directors in a less disturbing form than that of an outright purchase. Slack time of mechanics can be employed, for instance. The advantage, too, of being able to suit one's own needs may amply compensate for what may possibly be a limited standard of workmanship. As a matter of fact, there are comparatively few items that really need be made up in permanent form—much of the work can be more efficiently, economically and adaptably carried out with temporary " breadboards ". Then there is no grief if the specification has to be modified.

As regards commonly used general-purpose instruments such as oscilloscopes, valve voltmeters and audio oscillators, a compromise worth considering is the range of kits offered by certain firms who have carefully studied and tested their designs so that they can be easily made up into useful and reliable instruments at considerably lower cost than ready-make equivalents.

The gift of improvisation is a very valuable one. The experimenter who has to wait until he gets the correct material is not likely to go far. This is no argument for untidy, slipshod equipment. Most people, if not all, do better work with neat apparatus. But the writer has seen important research of the most refined accuracy being carried out in the box-room of a small suburban house with what looked like a collection of junk.

CHAPTER 2

Premises and Layout

2.1. CHOICE OF PREMISES

The private experimenter who perforce has to fit his " lab " somewhere into a house is apt to envy his professional brother who is allowed funds to build premises specially for the purpose. The latter, on the other hand, is apt to realize too late that his resplendent " Technical Wing " is abominably noisy, is riddled with interference, and is useless for giving a correct impression of domestic listening or viewing.

Although the requirements and facilities of private and professional workers are very diverse, it is hoped that the reader of the following discussion of the question of premises will find it possible, by means of slight mental adjustments, to adapt it to the scale of work in which he is interested. As it is the amateur who is most likely to be restricted in opportunities, it is he who is held chiefly in view, but if much of this chapter sounds rather poverty-stricken to some fortunate engineers there are always others whose material difficulties rival those of the private individual.

Assuming first of all that funds are available for special premises, an isolated building has much in its favour, provided that it is suitably constructed. This means substantial brick walls, if possible of the cavity type, and a lining throughout of some sort of building board. If there is a flat roof—of course it must never be *quite* flat—there should be an air space between it and the ceiling; and the outer surface should preferably be of an easily-cleaned white material, not a dull black which radiates heat in winter and absorbs it in summer.

The object of the lining is to help in keeping the place at an equable temperature; the huts sometimes hastily put up for technical staff cost a fortune to keep warm in winter, and are like ovens in summer. When a cold spell is followed by a sudden warm west wind, the resulting condensation indoors is ruinous. A building-board lining also provides reasonable acoustic conditions, and lends itself to serviceable decoration and to practical convenience generally. There are special lining materials for use where particular acoustic requirements have to be met.

If the technical department is to be extensive, it is generally far better that it should take the form of a number of small rooms rather than a very large one. Space can be better utilized if there are plenty of walls; and jobs can be carried out with less mutual disturbance. So an ordinary house of suitable size may be better than a special building.

2.2. HEATING AND LIGHTING

Heating is important. If the room is well heat-insulated, it is not extravagant to consider thermostatically controlled electric heating. The uniform temperature so obtained is very valuable for preserving accurate standards, to say nothing of personal comfort. Ideally, there should be complete air-conditioning. For the most precise frequency standards, the N.P.L. uses the cellars of a mansion, which are easier to keep at a uniform temperature than buildings above the surface. Whatever the heating system, it should not produce water vapour in the room as do certain gas and oil heated radiators.

Lighting is also important. Strong direct sunlight is not good for apparatus, but there is difficulty in getting enough all-north natural lighting without excessive window space and loss of heat. Double windows are considered faddy in this country, though it is difficult to see why.

It is difficult to have too much electric lighting. In addition to well-diffused general lighting there should be plenty of flexible arm lights, mounted clear of bench space, that can be brought right on to the job for reading meters and examining wiring and components in awkward corners. (Incidentally, where normal lighting fails to reveal internals sufficiently, a "pen" torch and dentist's mirror are a great help.) Fluorescent lighting may be all right, if the possibility of stroboscopic effects is remembered. If a tool or other appliance happens to be rotating at a synchronous speed it is made to appear stationary, with possibly unpleasant consequences to anyone touching it. Also a small percentage of fluorescent tubes have been found to cause radio interference. Even ordinary metal-filament lamps have been known to do so, and unless one is aware of such a possibility much time may be wasted in locating the source.

2.3. INTERFERENCE

The only perfect treatment of any sort of interference is to remove its cause. But that is not always practicable, so it is generally necessary to provide suppressors to prevent it from entering the laboratory. An isolated building lends itself to this, as a filter can be installed where the mains enter. Where complete absence of external fields is essential, it is necessary to screen the whole room with wire netting, which may conveniently be in the space behind the board lining. It is no easy matter ensuring the continuous effectiveness of room screening, especially at doors and other openings, and it is essential to filter all ingoing wires. Fig. 2.1 shows one solution to the problem.

For details of room screening and other anti-interference measures, see G. L. Stephens's *Radio Interference Suppression* (Iliffe).

2.4. THE AMATEUR'S LABORATORY

Let us now consider the amateur, who may have no choice whatever of premises, nor perhaps even the exclusive use of a room. Where the house is old and large, the chances of appropriating at least an attic or basement are quite good, but a modern house is generally just too small to hold the family. As regards choice, if any, between attic and basement: in the absence of further data the author is disposed to advise the latter, because a screened transmission line can be used to make connection with a distant aerial, whereas a long earth lead has disadvantages that are not so readily overcome. This may not apply quite so decidedly to a steel-framed building, which, electrically, is almost a continuation of the solid earth.

One may be obliged to consider a roof loft. It has been asserted that lofts make excellent workrooms. The author's personal views

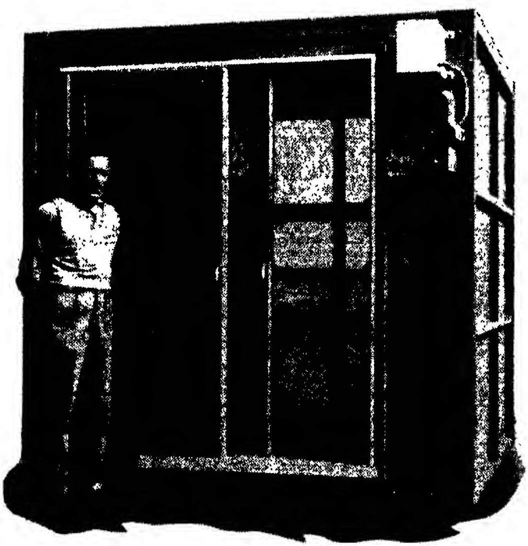


Fig. 2.1—A completely screened test cabin. Note the mains interference filter and the 1 : 1 mains transformer for reducing risk of shock by providing an unearthed power supply. (Belling & Lee, Ltd.)

do not incline in this direction, for his experience of them is that they can never for long be other than dusty; that they enjoy what the geography books call a Continental climate, being too cold in winter and too hot in summer; and that the trapdoor constitutes a grave danger to the preoccupied research worker. Even if the personal drawbacks are heroically overlooked, it cannot be denied that dust and extremes of temperature and humidity are particularly bad for apparatus. Nevertheless the author has seen a loft laboratory used successfully for really advanced work. Seclusion is an advantage that can hardly be overestimated, and one gets it better in a loft than in most places. An aerial lead-in can generally be arranged to come conveniently above the operating bench; but it is worth going to the trouble of ensuring that it is truly weather-tight. Another advantage of a loft is that one can string wires around, and generally damage the premises in ways that would never be tolerated downstairs—or should one say downladders?

Sometimes an outdoor shed can be appropriated, or built. Conditions here are not unlike those just described. In winter it is difficult to make it an attractive retreat, and damp weather has disastrous effects on the equipment, unless there is the unusual luxury of permanent heating. A concrete floor keeps the feet thoroughly cold in winter in spite of efforts to heat the place, and increases the danger of shock. A board floor is much to be preferred. In summer, too, both worker and apparatus may suffer from the climate in a hut or shed. But oppressive restrictions on the research worker are less likely to be enforced here than indoors.

2.5. STORAGE

If considerations of comfort, convenience, or necessity indicate an indoor room, and the experimenter is not in a position to exercise the rights of dictatorship, it may be possible to come to terms with other interests by carrying on the more unwelcome activities—such as construction of apparatus—out in the shed or garage. If the laboratory must, as a last resource, be shared, the guiding policy is to arrange for everything to be shut up under cover and out of sight when not in use. An old roll-top desk is an example of something that makes a good nucleus; the desk surfaces can be used for working, and the pigeon-holes and drawers for storing apparatus, etc. An old-fashioned wardrobe of ample proportions, with sliding shelves, makes an excellent store cupboard, and may even be adapted as a sort of work cabinet.

Whether or not a strict cover-up policy is forced on the experimenter by other members of the community, it is not a bad policy to adopt anyway. It is worth acquiring such cupboards, bureaux, filing cabinets, or nests of drawers as may be available, for storing instruments, components, tools, wire, valves, papers, and so forth. Good organization in this respect is worth while in saved time and space, creates a favourable impression, and keeps things in good order and condition. A chest of many shallow drawers is much better for storing the smaller articles than the usual deep drawers, which necessitate things being bundled on top of one another. Some drawers should be subdivided, eggbox fashion, to take valves, components, small tools, etc., in classified arrangement, so that it is not necessary to rake through the whole lot every time in order to find the right one. There is now an excellent selection of ready-made steel units for meeting any storage requirements.* Incidentally, it is a good thing to

* For frequently used small components, screws, etc., the units by Smallbone & Son Ltd., 116 Raddlebarn Road, Birmingham 29, are particularly convenient.

run through one's stock of components now and then, particularly of resistors and capacitors, to check the values. This may save much perplexity caused by taking them at their face values. And few activities repay the trouble so fully as descriptive labelling of all stocks and papers.

2.6. LAYOUT

Space is nearly always at a premium; and while it is very nice to have plenty of large tables or benches on which to set out one's work it is very seldom that there is room for this sort of layout. It is therefore necessary, as in the city of New York and for the same reason, to resort to vertical building. A very compact arrangement consists of a wooden cupboard placed, not in the usual position against the wall, but sticking out into the room, so that the back can be used for supporting instruments, switchboards, shelves, etc., above the work bench. A space of a few inches should be left behind the backboard for wiring.

If there is enough room, one may have further benches around the working-space, forming a sort of cubby-hole where everything is

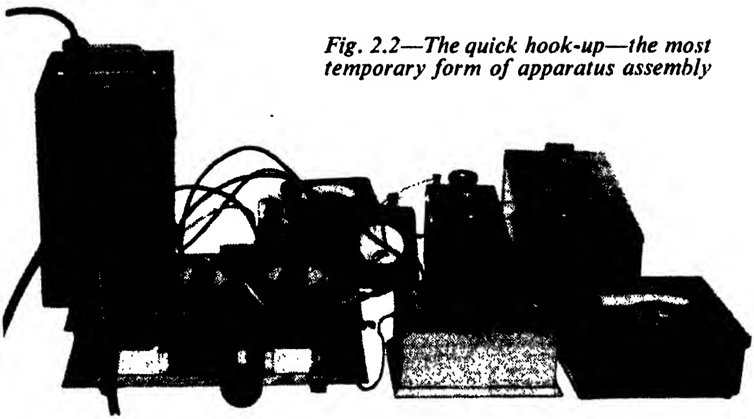


Fig. 2.2—The quick hook-up—the most temporary form of apparatus assembly

near at hand, except the less-used gear which is kept on the other side. At all costs the tendency to use bench-top space for storage must be combated. If it is not, the actual working-space soon becomes crowded out.

2.7. MOUNTING OF APPARATUS

Methods of setting up apparatus are worth careful consideration. The following types of layout are arranged in order of increasing permanence.

The Quick Hook-up (Fig. 2.2) consists of a number of separate components—valve holders, transformers, etc.—fitted with terminals and linked up by lengths of wire. This method is the most temporary of all, and, apart from students' experiments which have to be cleared away the same afternoon, has few justifiable applications. If there is to be any real saving of time the terminals must be adequate, preferably of the double- or triple-decker type, or a standardized system of plugs and sockets, and there must be an ample supply of prepared leads. High inter-wiring capacitance, poor contacts, and short circuits are probable if operating components are not firmly mounted.

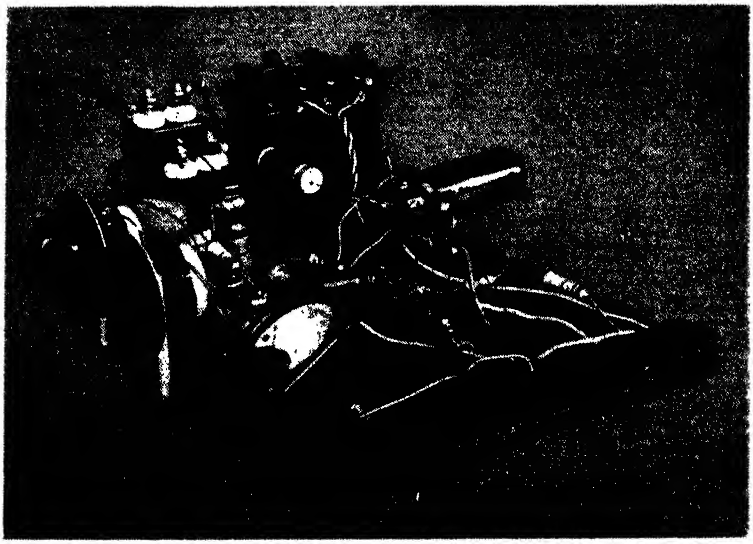


Fig. 2.3—An example of "breadboard" assembly

Fig. 2.4—Shrouded connector blocks are better than ordinary terminals for most purposes, temporary or permanent. This Belling-Lee type has the advantage of flexibility

The layout is non-portable, and if the experiment drags on it may be an encumbrance.

The Breadboard (Fig. 2.3) takes a little longer to set up and dismantle, and is not defintely a horizontal-space saver; but it is almost indispensable for assembling a circuit for temporary use, or for checking a design before irrevocably drilling holes in the permanent panel. The assembly forms a portable unit, and can be moved away to make room for other things. It is superlatively cheap, and often works much better for permanent use than more sightly units. But it collects dust. In the ordinary form it consists of a slab of wood, preferably ply not less than ½ in thick, on which are screwed all the components. One can easily shift components about and experiment with different wiring arrangements, and everything is accessible. Operating components—variable capacitors, potential dividers, etc.—are generally mounted on small vertical strips of wood screwed into the edge of the board; a stock of such components, ready-mounted, is useful. Terminals for external connections can be similarly mounted, but a safer and more compact form is the shrouded connector shown in Fig. 2.4, obtainable in strips that can be broken off into the desired number of " ways ". A piece of card placed underneath and screwed down with the strip so as to project is convenient for marking identifications.

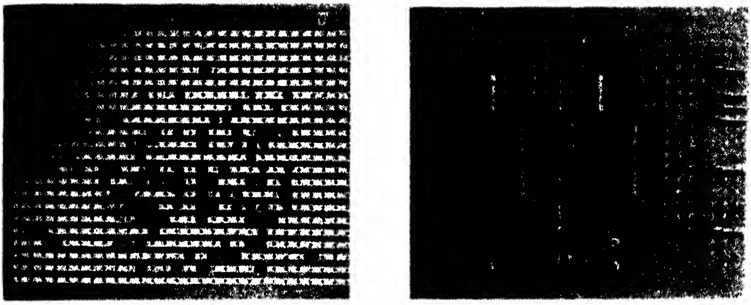


Fig. 2.5—" Veroboard " for temporary or permanent circuit construction. The rear (left) is provided with perforated copper strips which can be used for connecting the components seen on the front (right). Cross-connections between strips are made with pieces of wire pushed through the holes and soldered (Vero Electronics)

If the board has ½-in to ¾-in battens screwed along the edge on the lower side to form a hollow base, wiring can be more neatly carried out, and components are more easily mounted. Metal-coated plywood (Plymax) is convenient for use when a conducting base is needed.

An alternative to wood is thermoplastic material, to which anchorages can be made by pushing the heated ends of wires in and allowing them to set.*

Objections to the traditional form of breadboard are that for modern compact circuits it is clumsy, and that it bears little resemblance to the final form so that tests made on it are invalid. A convenient form of construction akin to printed wiring, free from these objections, and especially suitable for transistor circuits, is Veroboard.† Fig. 2.5 explains it.

The Chassis in its most conventional form is an inverted metal tray or shallow box, on the upper surface of which such things as valves and canned coils are mounted, while the smaller components and most of the wiring go below. Controls project from the vertical surfaces. In manufactured equipment the metal is usually cadmium-plated steel, but the mounts supplied by radio stores are almost invariably

* " Fabricating Circuits on Plastic Breadboards ", by J. H. Bigbee. *Electronics*, Sept. 1952.

† Vero Precision Engineering Ltd., South View Road, Southampton.

aluminium, which has the advantage of being easy to work, but not so easy for making sound electrical contacts. In the latter respect tin plate has much in its favour, and (if sloping sides are not objected to) a range of sizes is available in the multiple stores, where the product is better known as the baking tin.

The fact that this conventional arrangement protects the circuits from dust is an advantage in permanent equipment, but tends to be a disadvantage in experimental assemblies, where accessibility is paramount. Simply inverting the unit would be a help, if it had a more satisfactory means of support than the tops of the tallest cans. Moreover, drilling additional holes after the circuit has been wired presents difficulties and hazards. To avoid these disadvantages and to enable experimental circuits to be built up easily and quickly, several systems

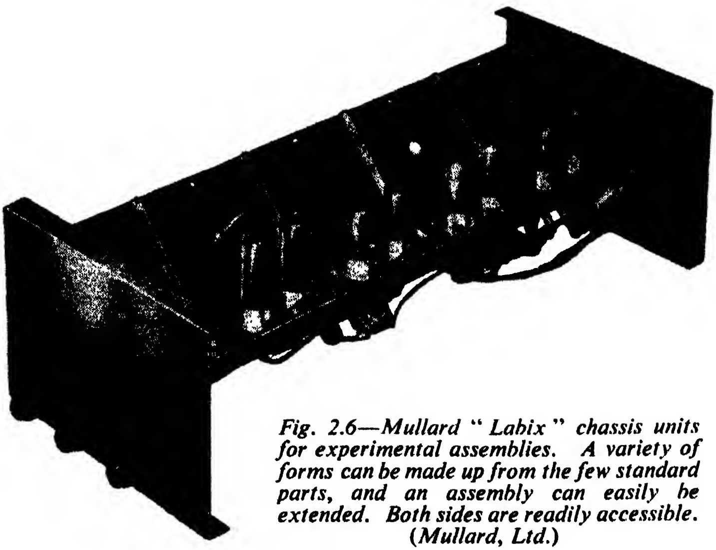


Fig. 2.6—Mullard " Labix " chassis units for experimental assemblies. A variety of forms can be made up from the few standard parts, and an assembly can easily be extended. Both sides are readily accessible. (Mullard, Ltd.)

of standard mounts have been put on the market. They can be used repeatedly, with resulting economy.

The " Labix " mounts* (Fig. 2.6) are designed primarily for experimental purposes, but are also suitable for permanent " one-off " equipments. They consist of channel shaped sections, with open middle portions to take small interchangeable panels for valves, etc., and end plates. In a rather similar system† the small panels are

* Mullard Ltd., Torrington Place, London, W.C.1. See also *E. & R.E.*, July 1957, p. 273.

† Cowell Developments, 67, Long Drive, London, W.3. See also *W.W.*, Oct. 1957, p. 487.

supported on perforated rails, to which brackets can be attached at right angles to the panels. A modification of this system, using only one standard metal part, supplemented by Meccano angle girders and wooden plates, has been described.*

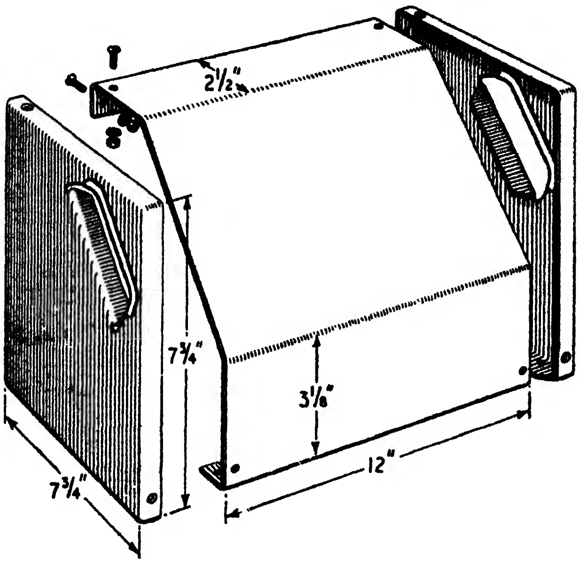


Fig. 2.7—Sketch showing construction (with leading dimensions) of the B.B.C. standardized chassis. The sloping panel is supported at the sides by two flanges bent in from the cut-away slots that form the hand grips

Another form of chassis, useful for both temporary and permanent work, is that devised by B.B.C. engineers (Fig. 2.7); it consists of a pair of cast aluminium end plates supporting a sheet panel. The end plates can be used over and over again as panels fall out of use.

A versatile commercially obtainable† chassis construction system is the "Lektrokit," the basic components of which are perforated aluminium-alloy chassis plates, rails and end plates. With the addition of front panel and covers from the kit, a temporary rig can be put into respectable dress for permanent use.

The conventional chassis lends itself to incorporation in boxes and cabinets of almost any type. One particularly useful development is the Widney-Dorlec‡ system, in which a framework is made up to the required size from standard sections, cut to length, and die-cast corner

* "Prefabricated Chassis", by D. M. Neale. *W.W.*, Aug. 1957.
† All-Power Transformers Ltd., Chertsey Road, Byfleet, Surrey.
‡ Hallam, Sleigh & Cheston, Ltd., Bagot Street, Birmingham, 4.

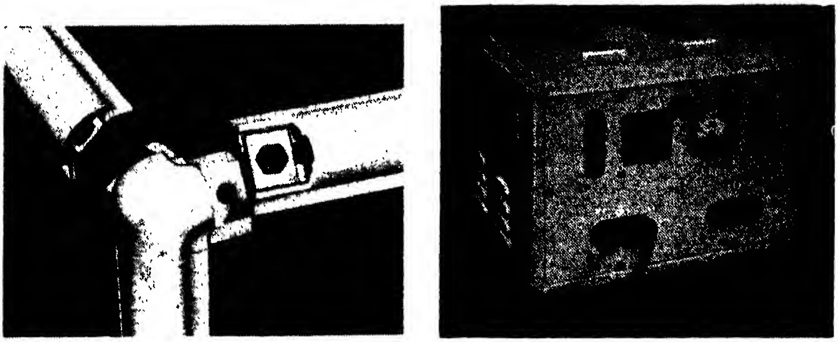


Fig. 2.8—Example of Widney-Dorlec construction, showing (left) corner pieces and sections and (right) appearance with typical instrument in position. (Hallam, Sleigh & Cheston, Ltd.)

pieces, as shown in Fig. 2.8. Although easy to construct, the result is neat and professional.

The Platform Frame (Fig. 2.9) comes into being when several units of the breadboard or chassis type are to be combined, or when a more compact arrangement than a single large board is desired. The frame and platforms may exist as a self-contained structure, on which any assemblies can be supported temporarily; or the breadboards or chassis may themselves form the platforms—an arrangement more suited for permanent assemblies. In either case, the construction ought to permit any one unit to be removed for inspection. If desired, the sides can be filled in with panels, dust covers, or protective wire-gauze screens. Or a popular type of tray filing cabinet may be adapted as the structure, provided that ventilation is introduced where necessary. This system is a cheap and flexible one, especially suited to amateur requirements and capabilities. It used to be the most popular arrangement for senders, but is now giving place to

The Vertical, or Telephone, Rack (Fig. 2.10). A substantial frame, usually of iron or steel angle or channel section 19 in. or 14 in. wide, is used to support vertical panels to any convenient height. It is obviously the best scheme when the apparatus includes many controls, meters, jacks, and so on; but almost any unit can be accommodated—for instance a broadcast-receiver chassis or a gramophone turntable. The most usual form of unit incorporating valves, etc., that cannot conveniently be mounted on the panel itself, is a ⊣-shaped structure consisting of a panel with a platform at right angles behind it. Bracing supports are needed when the unit is heavy and projects far back or forward. Dust covers removable at the back protect the apparatus, and the wiring between units may if necessary be led through screening conduits. The system is rather expensive to carry out properly, but makes a very neat and adaptable job for permanent use. A modification, valuable for accessibility, is to hinge the panels,

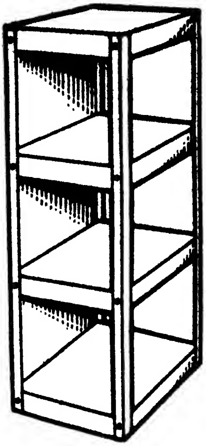
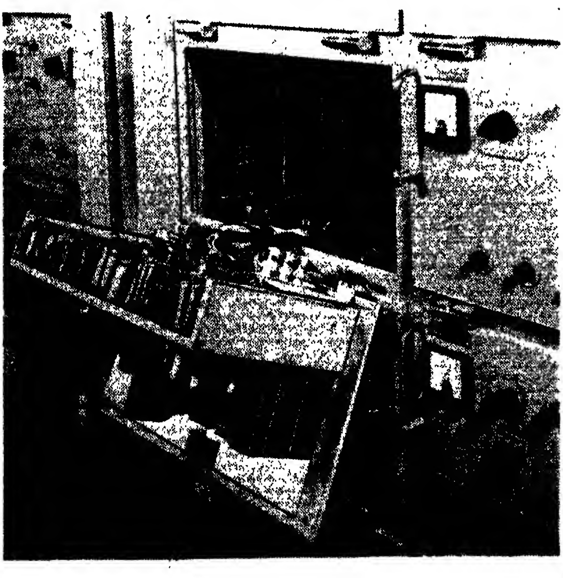


Fig. 2.9 (left)—Platform frame and rack for mounting several units in tiers

Fig. 2.10 (right)—Vertical rack mounting for panels comprising a large assembly

Fig. 2.11—Rack-mounted panel hinged for accessibility



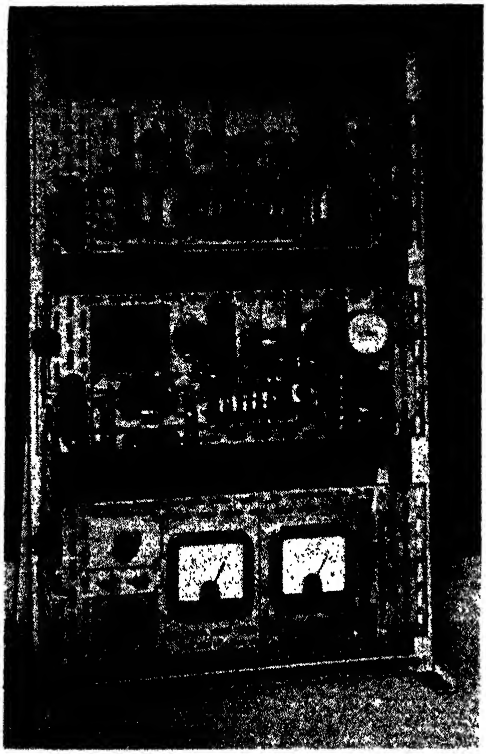


Fig. 2.12 (left)—A space-saving and quickly constructed framework made up from "Rapikon" standard parts. It provides good accessibility and easy modification and extension (Shandon, Ltd.)

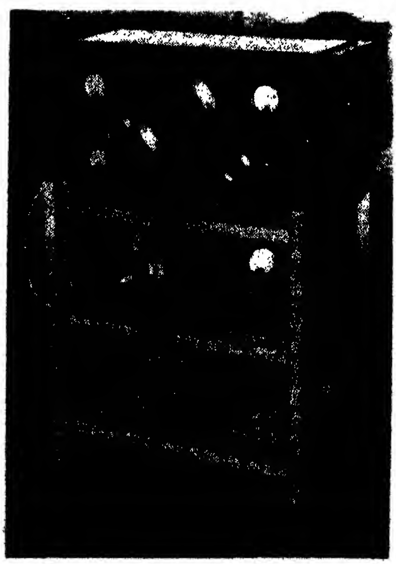
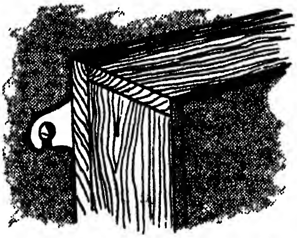


Fig. 2.13 (right)—Trolley mounting is particularly suitable for much-used bulky units

Fig. 2.14 (right)—Oscilloscope mounting trolley (Nagard, Ltd.)

Fig. 2.15 (below) — Mirror plates screwed to box-form units enable them to be mounted on vertical surfaces



as shown in Fig. 2.11. Another variation, favoured by one reader on account of accessibility, is to mount the chassis upside-down, but the disposal of valve heat must be considered.

For experimental or semi-permanent vertical mounting, the " Rapikon " standard parts* (Fig. 2.12) save much work.

An inexpensive material, useful for making up racks as required, as well as benches and much other laboratory construction—a sort of life-size Meccano—is Dexion Angle.†

The Trolley (Fig. 2.13). Sometimes one has a bulky unit that occupies valuable space in any one position, and is difficult to lug about or apply at the most useful points. The solution is to mount it on a trolley. It is a particularly convenient receptacle for such things as cathode-ray-tube apparatus. Fig. 2.14 shows a specialized form of mobile mounting for oscilloscope and power unit.

Instruments as bought, and sometimes as made, take none of the aforementioned forms, but consist essentially of boxes. It is a good idea to fit these with mirror plates to enable them to be hung one above another on the wall or backboard of the bench (Fig. 2.15).

For more information on general design of equipment for amateur and small laboratories, see an article by W. D. Cussim (*W.W.*, March 1958).

A most admirable and detailed account, with coloured temperature diagrams, of research into the problem of cooling electronic

* Shandon Electronics Ltd., 6 Cromwell Place, London, S.W.7. See also " A New Electronic Assembly System ", by H. C. Bertoya. *E.E.*, Feb. 1958.

† Dexion Ltd., 67, Maygrove Road, London, N.W.6.

equipment, has been given by E. N. Shaw (*E.E.*, Jan., Feb., and March 1957).

2.8. BENCHES AND WIRING

If it is decided to construct a work bench, it is wise to see that it is strong and rigid. Boards are apt to shrink and warp, leaving crevices for small screws to hide in, so a sheet of thick linoleum or hardboard over the top is very pleasant for working. A shelf near the ground helps to brace the bench, and is most useful for supporting batteries, power units, and other heavy equipment. The usefulness of Dexion for the framework has already been mentioned.

Then there is wiring. An abundance of electric supply points should be provided; otherwise, when there is a soldering iron, a check receiver,

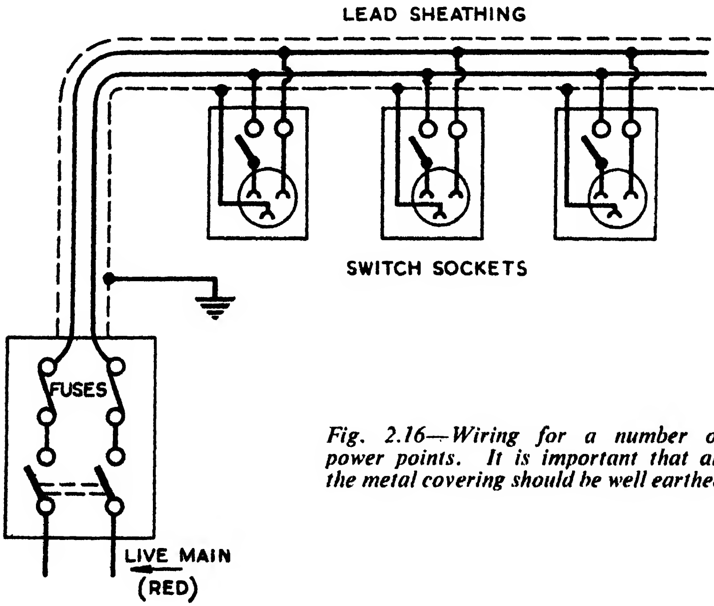


Fig. 2.16—Wiring for a number of power points. It is important that all the metal covering should be well earthed

an oscillator, and a voltage-regulated power unit going, it may be difficult to find another point to connect a bench lamp, a heater, or a fan. Thus it is a good policy to scatter sockets about fairly freely, each separately switched in the live side (Fig. 2.16). A good place is just underneath the top surface of the bench.

And here a warning must be sounded. There have been some sad examples of what the amateur and even so-called professional wireman can do: bits of ordinary wire twisted around nails, and so forth. There are several reasons for making a proper job of it. One is that shoddy wiring is a cause of fires. Another is that it leads to leakages which introduce hum and other undesirable effects, shock, and breakdown.

So use lead-covered electric-lighting cable of good quality, with all sheathing bonded together and soundly earthed—preferably by a different connection from that used for radio or other circuits. P.V.C. covered cable is usually used nowadays, but provides no screening. Probably the whole lot of sockets will have to connect to the mains at some one point in the room. Try to arrange matters so that, if a fuse goes because of some work being done, it does not plunge the place into darkness. For example, if the bench connections *must* be taken from the branch lighting circuit, provide fuses on the bench sub-branch of a slightly lower current rating than those protecting the whole branch. Incidentally, any mistrust with which the supply authorities may regard the system will be greatly intensified if they find lamps being used on a power circuit, at a lower tariff.

The type of plug which, at the time of writing, is a British Standard, contains its own branch fuse, which should be of a rating to suit the appliance to which the plug is wired. Plug connectors also have a third pin, for earthing. Bayonet sockets lack this, but for two-wire low-power connections the plugs are much cheaper and are less likely to fall out of under-bench sockets than the pin type.

A soldering iron is nearly always wanted at a moment's notice, so provision for heating it up quickly, and then keeping it on permanently at a low enough temperature to avoid burning, is valuable. Some stands incorporate a switch to cut in a resistance for stand-by; alternatively, a tapped auto-transformer or a Variac (Sec. 4.2) is very convenient for adjusting the temperature to working requirements.

Talking about soldering irons—old timers who are just beginning to deal with transistors and miniature components should be warned not to attempt to use the same substantial iron that has served them so well in the past. A special miniature iron should be used, and the wire lead held firmly with a pair of pliers between the iron and the component so that heat flowing from one to the other is shunted away. Unless the iron is a low-voltage type run via a transformer, care should be taken that the insulation at working temperature is sufficient not to pass an appreciable reverse current to earth through a transistor, risking its ruin.

2.9. EARTHING

Good screening and earthing are needed in any case, but especially where the supply is d.c., because the high-pitched commutator ripple has a way of getting mixed with the input of amplifiers, and is particularly unhelpful in bridge work.

For earthing, a main water pipe, close to where it goes underground, is generally satisfactory, and saves a lot of manual labour. But if a specially-made earth is needed its requirements are (1) plenty of exposed metal surface in contact with the ground, (2) burial deep enough to be in permanently moist soil, (3) no corrosion (lead tape or piping brought right into the room is good), and (4) a short run from ground to terminal. It is an excellent idea to bury the metal in two equal lots, brought out to separate terminals. Their condition can then be checked at any time by measuring the resistance from one to the other, preferably with a.c. to avoid electrolytic effects. The resistance of both in parallel cannot exceed a quarter of this figure. Another advantage of having two earths is that a single common one sometimes causes undesirable coupling between units connected to it.

2.10. CONNECTORS

An interconnection system is valuable if for any reason it is necessary to split up the lab, even if it is only a matter of two benches in different

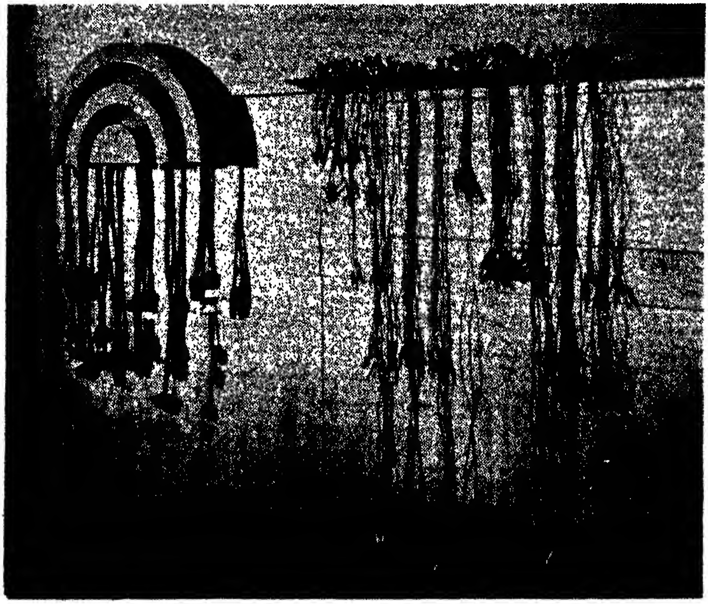


Fig. 2.17—How to store leads so that the right one can be selected without loss of time and temper (N.V. Philips Gloeilampenfabrieken)

parts of a room. Several wires, preferably of really heavy gauge, joining such benches and brought out to terminal strips or sockets, enable these parts to be used as a whole. If one space is fully occupied by the source of a test signal, for example, the signal can be piped through this line and used elsewhere. Or power supplies from a common source can be distributed.

Lastly; one must not fail to lay in a good stock of flexible connecting leads, single and twin, of various lengths from a few inches up to a yard or two; some with tag or tinned ends, some with crocodile clips, and others with suitable plugs and sockets. The use of different colours, especially for twin or multiple leads, saves time and guards against mistakes. Fig. 2.17 shows how to store them so that the right type and length can be selected without any trouble.

CHAPTER 3

Fundamental Principles of Measurement

3.1. MEASUREMENT V. GUESSWORK

The scene is a police court. The case is one of alleged dangerous driving. Witnesses are called. Some say the accused was travelling at 29 miles per hour; some 40; some 60. Some merely declare that he was travelling at a terrific speed. All very inconclusive and unsatisfactory. Personal judgment, ignorance, prejudice, rather than facts, determine the evidence.

Another case: the witness this time is a mobile constable, who states that when he was driving behind accused at the same speed his speedometer read 38 miles per hour. The inference, presumably, is that this is the speed at which the accused was driving. The data in this case are far more satisfactory; but not beyond challenge, as any good defending counsel would demonstrate. There would be penetrating cross-examination on how and when the police-car speedometer was checked, and how the witness knew he was travelling at the same speed. Translating it from the language of the law court to that of the laboratory, what were the instrumental and personal errors? In such an important matter as legal action, affecting perhaps the liberty and reputation of a citizen, evidence ought to include not only the figures but also the limits of probable or possible error. At present that seems rather too much to expect of a law court, but it is accepted practice in a laboratory. Measurement is the basis of scientific progress; opinion, guesswork and assumption are its chief enemies.

3.2. FALSE ASSUMPTIONS

One might suppose that in the very Temple of Science itself, the laboratory, there would be no room for these reactionary forces. But they are constantly at hand, ready to obscure one's work. Radio itself would probably have been established earlier if authorities had not decided that it was impossible. Later, investigators noticed that as the wavelength of transmission was reduced the range apparently became less and less; so they assumed that very short waves would be useless, and allocated them to amateurs. Subsequently they found that they had been more generous than they had intended, when it turned out that these waves were the *only* ones suitable for very long ranges.

How often when carrying out experiments one observes that y steadily increases (or falls) in some sort of proportion to x, and is tempted to miss out the last few readings and extend the curve by freehand. Or when some unexpected effect is noticed, how easy it is to make an explanation for it that will save the trouble of really finding out. It is nearly always worth while following up anything that does not work as it should, rather than dismissing it as "experimental error", or cooking the results until they come "right". It may be the key to something new. At the worst it will clear the matter up and give one a better grasp of the subject. The difficulty is that in these days work must be got through so quickly, and there is no time for exploring every little side track that appears in the course of routine. Even if there were time, few would have patience to check the obvious. But it is the ideal to pursue.

3.3 DIRECT OBSERVATION OR INSTRUMENT READINGS?

Within the field of electrical communication and electronics there are very many different things to be measured and tested, and there are usually several possible methods of dealing with each. The variety of equipment for these methods is quite bewildering. So instead of plunging at once into an ocean of detail it would probably be helpful to take a broad view, and consider the principles governing choice of method. It must be remembered, however, that broad general principles cannot be expected to apply rigidly to every particular situation.

It is generally taken for granted that accuracy is the main object. The quest for accuracy is certainly the motive of most of what is written in this and other books on laboratory work, but mere percentages are not enough. The first thing to ask about the proposed method is whether it is going to tell us what we ought to know. It is not always realized that the ultimate object of our work is to influence the human mind, through the senses. The particular thing under examination at the moment may only be a resistor, but the purpose of examining it is, say, to enable it to contribute to the better working of a television receiver, which makes its influence on the mind through the senses of seeing and hearing. In the end it will be judged by the mind. So the mind is really the ultimate measuring instrument. But unfortunately it is a most unsatisfactory instrument for arriving at definite conclusions. A listener may say that loudspeaker A sounds better than B. But he is unable to say *how much* better except in the most general terms. Another listener may give quite a different estimate, or even deny that it is better at all. And any one listener's estimate may vary according to whether he happens to be alert or tired, pleased or irritated.

It might seem that a direct switch-over between the two things being compared would allow a fair and easy decision. But long experience of such tests shows that this is not so. A judgment made on one occasion may be reversed later when the programme changes. The indefiniteness of such work makes performing it for any length of time curiously tiring and unsatisfying.*

The alternative is to measure the physical quantities that produce the mental sensations. For example, instruments can be used to measure the strength and waveform of a sound, or (a step farther back still) of the electric currents generating the sound. While this certainly makes it possible to obtain precise and consistent data, it raises a new difficulty. There seems to be no logical correlation between the physical thing and the human sensation. A person born stone deaf can learn all about the physics of sound, but this does not help him in the least to know the sensations produced by it; nor can other people's explanations convey it to him. True, we learn by experience that the greater the sound pressure the greater the sensation that we call loudness, but the two do not seem to be connected by any simple law. There is not even any precise method of measuring loudness. When sound is distorted, or disturbed by noise, the resulting sensation is generally unpleasant. But nobody has succeeded in establishing a definite relationship between the amount of distortion or noise and the amount of unpleasantness†. This being so, measurements can be quite misleading. On the face of it, a reading of 20 milliwatts of noise on an output meter is evidence of more noise than one of 5 milliwatts. But 5 mW of one kind of noise may be found to be more disturbing than 20 mW of another kind.

So the first step is to decide what one is really trying to discover. If it is something that in the end will be judged by the mind, one is faced with the dilemma that a personal (subjective) test is unreliable and indefinite, while a physical or instrument (objective) test has to be interpreted, and the interpretation is open to dispute.

3.4. SUBJECTIVE METHODS

The large organizations that study these matters, having realized the unreliability of individual judgment and the difficulties of interpreting physical tests, have developed a technique based on the principle that random errors tend to cancel out when the average is taken of a large number. To preserve the desirable subjective basis of the experiment, personal judgment is used, but the independent impressions of a large number of people are systematically taken and combined. In addition to giving an average result that is more representative than a single one, this process indicates how far individual results spread around it. The persons are carefully selected to represent the public generally, or to be all untrained, or all trained in observation, or on some other basis, as the experiment may require. The whole thing is in the highest degree systematic, and far from being a mere collection of opinions.

* This has been found to apply even to pure tone tests; see a paper by A. Stott and P. E. Axon, *Proc. I.E.E.*, Part B, Sept. 1955.

† E. R. Wigan (*E.T.*, April and May 1961) has made one of the most thoroughgoing attempts, but would hardly claim complete and unquestionable correlation

Subjective methods are unlikely to be very conclusive unless properly organized on a large scale. So they are outside the range of most experimenters. For the sake of definiteness and the ability to compare results obtained by different people and at different times and places, the choice nearly always falls on objective methods. When these are in lieu of subjective observation, the problem of interpretation arises. It is a fascinating and controversial subject, with plenty of scope for further research. And it is important. The most refined measurements may be misleading if wrongly interpreted. But the only place that can be given to it in this book is the warning to take great care in drawing conclusions from instrument readings about subjective things, such as loudness, brightness, and intelligibility.

Much lab. work, however, might be described as sub-contracts, not directly concerned with the ultimate purpose. Such jobs, for example, as measuring resistances and inductances, obtaining valve characteristics, and determining selectivity; in these only the physical facts are needed.

3.5. DISENTANGLING THE CONCLUSIONS

In any case, the subject from now on will be confined to physical terms, and the question will be a choice from objective methods. Even then it is necessary to make sure that the right conclusions will be drawn from what the instruments tell. To take a very simple example: the books say that if you interpose a capacitor of, say, 100 pF in series with a large aerial you improve the selectivity, at the cost of reduced signal strength. You try it, and the signal strength goes up. This does not prove that all the books are wrong. It just shows that in this particular receiver the large aerial was too tightly coupled, and so overdamped and probably mistuned the circuit; or else that the receiver tended to be unstable due to stray feedback, which was kept in check by the large aerial and released by interposing the capacitor. Or perhaps a bit of both. In any case, the effect quite correctly predicted by the books was concealed by a larger effect due to the peculiarities of this receiver, and unless the experiment were rearranged it might be impossible to disentangle the required result. It would be no good contributing a paper to a scientific society, claiming to have proved by means of experimental evidence that the accepted theory with regard to aerial series capacitors was wrong. The experiment was not conducted with sufficient attention to the possibilities of irrelevant influences acting at the same time.

So make sure of one thing at a time. If the quantity being investigated, x, depends on factors, p, q, r, s, and t, and it is desired to find how it depends on p, make sure that while p is being varied the factors, q, r, s, and t remain constant. Otherwise there is confusion.

The objective having been determined, the next concern is to achieve it with the greatest accuracy. That is the ideal, but it may have to be trimmed a little to circumstances. Sometimes it is more important to obtain results quickly or easily or inexpensively than to spend much time, effort, or money on greater accuracy than is really needed. The sin of using more resources on a job than are necessary is second only to that of using insufficient to do it properly. And obviously one's choice of method is always subject to apparatus being available—or capable of being extemporized. These qualifications having been made, the remainder of this chapter will have accuracy as the motive.

3.6. ACCURACY AND PRECISION

And here it may be as well to distinguish between accuracy and precision. Like many other words, such as work and force, scientific usage endows them with a more limited and precise meaning than they bear for general purposes. The distinction can perhaps best be shown by an example. A certain cheap clock fitted with a seconds hand can be read quite easily to the nearest second, but owing to temperature and other influences its error 24 hours after resetting may be anything up to several minutes. Its precision is quite high but its accuracy poor. Except over short time intervals, its precision could be considerably less without appreciably affecting the overall accuracy. Big Ben, on the other hand, can normally be relied upon within one second, but when viewed from the street cannot be read nearly so precisely. So in these circumstances much of its accuracy is lost. Considered another way, its overall accuracy is determined almost entirely by the reading error, or lack of precision. Precision is necessary for high accuracy, but is not sufficient for it.

Another important thing to realize is that it is not enough to say that the accuracy of an instrument is so-much per cent. For one thing, it would obviously be more sensible to call it the *in*accuracy, and then there would be no doubt about whether a greater accuracy meant a greater percentage or a smaller one. But other questions arise. Does the figure given mean the greatest possible difference between the value shown by the instrument and the absolute value at the time of calibration; or does it mean the greatest subsequent variation in value—and, if so, over what range of temperature, humidity, frequency, etc.? The latter by itself can be distinguished by the term *constancy*. (See Sec. 5.3 for a simple example of instrument inaccuracy.) The subject is discussed at length by W. H. F. Griffiths (*W.E.*, March 1943).

3.7. BASIC METHODS: DEAD RECKONING

The innumerable separate methods of measurement can be classified into a few basic types; and doing so helps one to see the fundamental principles more clearly and to judge which of several methods is likely to be best for the purpose in view.

First of all we distinguish between passive quantities or parameters (such as resistance, capacitance, and attenuation) and active quantities (such as voltage, current and power). Passive equipment includes resistors, capacitors, and attenuators; active equipment, such items as batteries, power units, and oscillators. In general, passive gear is more constant and reliable, and results obtained in terms of passive quantities are more likely to be accurate.

When the object being investigated is active—for example, a battery—it is usually possible to employ its own activity to provide the answer. The method is to apply some sort of indicator, which in this case might be a voltmeter. A high grade of instrument-making is necessary to ensure accuracy of a better order than 1%. And the power needed to work the indicator is often a substantial part of that available, so the quantity to be measured is liable to be altered merely by connecting the indicator. This is the simplest class of measurement, illustrated by the first of the simple functional diagrams in Fig. 3.1. In all of these, X denotes the object being investigated and I the indicator.

If X is passive there is nothing for an indicator to respond to, and it is necessary to provide a source, S, as at *b*. For example, the resistance of a piece of wire (X) can be measured by applying a battery (S) of known voltage and using an ammeter or milliammeter (I) to read the current passed by the wire, from which information the resistance can be calculated. Observe that the voltage of the battery must be assumed or measured, and the meter accurately calibrated over a suitable range. An error in either of these affects the result.

In both *a* and *b* the measurement is indirect, in the sense that the unknown is deduced from quantities of a different kind. In comparison with the methods still to be noted, however, these two methods may seem to be direct, in the sense of being straightforward. So to avoid misunderstanding they will be classified as *dead-reckoning* methods.

3.8. SUBSTITUTION

The next two are classed as *substitution* methods, because X is measured by putting a known standard of the same kind (labelled C in the diagrams) in its place, for comparison. If C is variable over a range that includes the value of X, there is no need for I to be calibrated at all; all it has to do is to enable C to be adjusted to give the same indication, after it has been substituted for X. Even if C is not continuously variable, I may not have to be calibrated in actual units, so long as its readings are in strict proportion.

Method *d* has the additional and very important advantage that the standard is passive. Applied to measure the resistance of the piece of wire, C would be a standard resistance box. Any battery and current indicator could be used, so long as a reasonable deflection was given. When the deflection of I had been noted as in *b*, S and I would be switched over to C (as shown dotted), and C adjusted until I gave the same deflection. The resistance of C would then be the same as

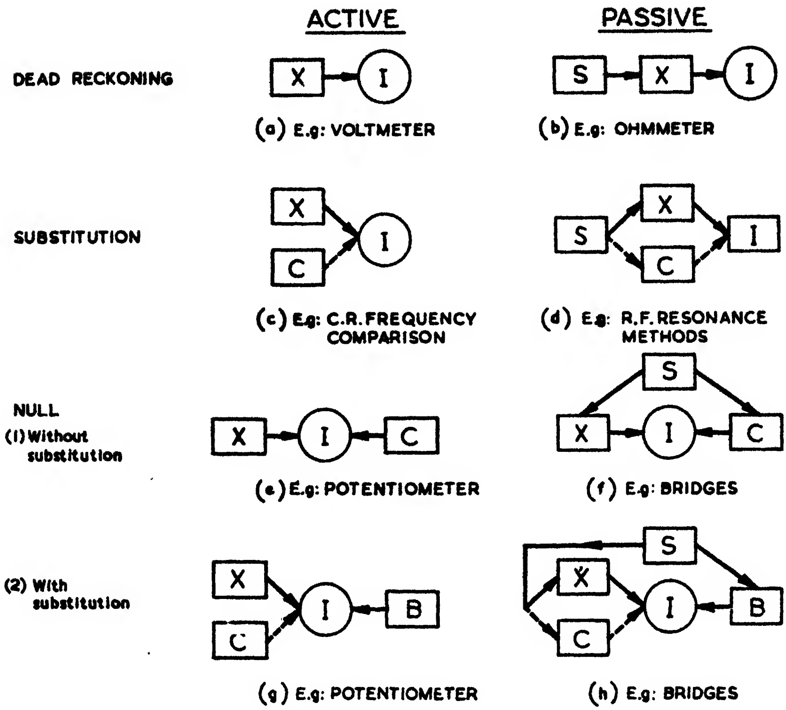


Fig. 3.1—Functional diagrams of the basic methods of measurement. X = unknown—to be measured; I = indicator; S = source of power or signal; C = standard of comparison; B = balancer

that of X. Here, then, we have the answer in terms of one datum only, of the same kind as the unknown, and passive—all conditions that make it easy to obtain high accuracy.

Incidentally, the next three chapters, on equipment in detail, deal respectively with what are here denoted by S, I and C.

3.9. DIFFERENCE MEASUREMENTS

Considered in this broad theoretical sort of way, the method just described may not seem to leave much room for further elimination of causes of error. But however pleasant it is to think of a standard as a lump of the required quantity—resistance, inductance, or whatever it may be—to the value indicated by the dial reading (and diagrams certainly tend to encourage the habit), this view does not correspond with reality. Electrical quantities are not confined within clearly marked boundaries; they are distributed throughout and around the circuit. One must not leave out of account such humble but essential features as connections and switches. So the comparison in Fig. 3.1*d* is only theoretically between an accurately-known standard and the unknown; in practice there are sundry uncertainties that more or less upset the comparison. When measuring low values of resistance, for example, the resistances of leads and switch contacts may be appreciable and not necessarily the same in both positions. The standard resistance may be known to within 0·1 %, but this is of little advantage if incidental circuit resistances amount to perhaps 2% and are not known at all. Comparison of small capacitances is even more liable to be vitiated by stray quantities.

Complications of this sort can usually be wholly or largely excluded from the problem by dealing, not in absolute quantities, but in differences. To follow what this means, suppose the standard consists of a calibrated variable capacitor, and when used as in *d* it reads 37·6 pF. But how can one be sure that the stray circuit capacitances, which can hardly be less than several pF, are exactly the same as when the instrument was calibrated? The answer, usually, is that one cannot. So the figure 37·6, which is supposed to be the value of capacitance, is more or less uncertain. In any well-arranged standard, however, this uncertainty affects every reading equally. So, by taking the *difference* between two scale readings as the standard, the uncertainty cancels out.

The modification of *d* needed to use the standard in this more reliable manner is to have it in circuit all the time. First it is set to some high value, such as maximum. Then the unknown capacitance is inserted and the standard reduced until the total is the same as before. The unknown is given by the difference between the two readings on the standard. The same stray capacitances are in circuit all the time, so do not upset the comparison. Similarly, when comparing resistances the standard box is kept in circuit all the time, so that as much as possible of the circuit is the same for both readings, the alterations being confined as far as possible to those parts that are meant to be compared.

So whereas the absolute value of a standard is generally more or less uncertain because one may not be sure how far it takes account of end connections, stray capacitances, and so forth, a standard can be calibrated in *differences* with certainty. The difference method of comparison is therefore to be preferred.

3.10. NULL METHODS

There is still room for improvement. The accuracy of all the foregoing methods is ultimately limited by the precision with which the indicator can be read. If the scale is 4 in. long and can be read to 0·01 in., at one-quarter full scale the reading error may be as much as 1%. It can be reduced to 0·25% by increasing the power of the source or the sensitiveness of the indicator to bring the deflection up to full scale. But any further increase just drives the pointer off the scale. However, if the method is so arranged that the comparison is made by adjusting the indicator reading to zero, there is hardly any limit to the sensitiveness of the indicator that can be used. So the reading error can be made as small as one likes, and full benefit obtained from the accuracy of a high-grade standard. But of course if the accuracy of the standard is relatively low, more precise comparison can do no more than give one an over-optimistic idea of the accuracy.

Varieties of this, the *null* method, are represented by the remaining functional diagrams in Fig. 3.1. In *e*, the active unknown is balanced against an active standard in such a way that when they are equal the indication is zero. An example of this is the measurement of potential difference by means of a potentiometer and standard cell. The corresponding passive scheme is *f*, and this is particularly important because it includes many bridge methods. It should be noted that C is not necessarily of the same kind as X. A valuable feature of *e* and *f* is that even though the indicator is made to see the effects of X and C as opposite and equal, the comparing system can be so devised as to introduce a known multiplying ratio, so that X can be measured over a much larger range than the available values of C. But of course this system itself introduces some error, and it may be difficult to allow for the various strays affecting C and X. So for the highest accuracy, or at any rate the least uncertainty, the comparison should be made by direct substitution, as in *c* and *d*. To retain the null indication, a balancing element, denoted by B in *g* and *h*, must be used to cancel out the indication received via X and C. There is no need for the value of B to be known, but it should be very steadily and critically variable. Method *g* is not really important, but is included to make the table complete; *h* represents another class of bridge methods. All these methods can be subdivided into the absolute (switch-over) and difference varieties, described in connection with *c* and *d*.

3.11. THE QUEST FOR ACCURACY

Broadly speaking, then—which is all we are doing just now—the foregoing analysis of methods is in order of increasing accuracy. The fewer data, such as calibrations, that have to be taken for granted the better. And the simpler and more fundamental the standard, the more likely is the result to be accurate. An example of a type of measurement offering the possibility of very good accuracy is one in which the unknown is compared (by a null method, to minimize reading errors) with a single stable passive standard, using a difference calibration. In making a complicated experiment the bench may be covered with instruments, and if each of these is subject to an error that affects the result the final accuracy may be very poor. The experienced worker devises his method so that the factors affecting the result are confined to a few that are well under control, preferably the most accurately known of those available; and he arranges for unavoidable errors to cancel out in the result.

Take for example a valve voltmeter. One way of using it to measure r.f. resistance (Sec. 9.30) necessitates taking two or more readings of voltage, and also relies on a similar number of known r.f. resistances, probably measured at quite a different frequency. Apart from that, the calibration of a valve voltmeter of the usual type depends on a number of factors that may not be known to great accuracy, and may have altered to an indeterminate extent since the instrument was last checked. The accuracy of this method depends on widely different voltage readings, yet the more widely they differ the less likely are the voltmeter errors to cancel out when deriving the result. But if the method is devised so that the voltmeter is used merely to indicate equality in two tests, the most that one assumes about it is that the deflection due to that one particular voltage has not drifted during the short interval between the two tests. The calibration is not used at all.

The aim, then, is to work the method around so that the factors controlling the result are the fewest and most reliable in the laboratory, and so that the comparison itself does not introduce needless error. Every reading is subject to error, but some are much less so than others. So the best method depends largely on the relative reliability of the instruments available.

3.12. PERSONAL ERRORS

You yourself, the observer, are one of those instruments, and (with all due respect) perhaps not among the most reliable. It is extraordinarily easy to make slips in reading scales, noting the positions of range switches, connections, and conditions generally, and in performing calculations and drawing conclusions. So another aim should be to exclude these possibilities, by avoiding awkward scale factors, and not trying to do mental arithmetic at the same time as the observations. Do not hurry over the planning, and check the set-up carefully.

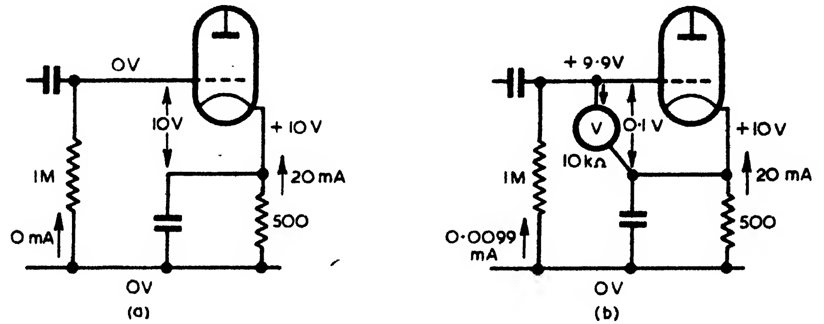
Provided that it does not restrict one's experience, it is advisable to keep to methods that are thoroughly understood and instruments whose capabilities and peculiarities are known.

Some notes on planning and carrying out experiments are given in Chapter 8, and on working out the results, in Chapter 13.

3.13. DISTURBING EFFECTS OF INSTRUMENTS

The need for avoiding irrelevant influences has been mentioned. A particular and important sort of irrelevant influence is the testing apparatus itself. You may wish to know the negative bias voltage on the grid of a valve under the working conditions specified in Fig. 3.2*a*. The simplest way, one might think, is to connect an accurate and reliable voltmeter to the grid and cathode, between which points the voltage is to be found. Certainly the voltmeter may give a reading. But it is wrong to assume that that reading is what is wanted. You want to know the voltage under working conditions. What you actually read is the voltage with a voltmeter—a conducting path—between the points. And the two things may be vastly different, as may be seen by comparing *b* with *a*. A better approximation can be obtained by measuring the voltage across the comparatively low-resistance bias resistor (*c*); but then you are not *quite* sure that the voltage measured is actually getting to the grid. There may be a hidden break in the grid leak, or some leakage from grid to cathode or through the coupling capacitor. A still better method is to note the voltage from some independent supply that can be connected from grid to cathode without altering the anode current.

Or consider measuring the gain of an amplifier. A switch is used to connect a meter to measure in turn the input and output voltages.



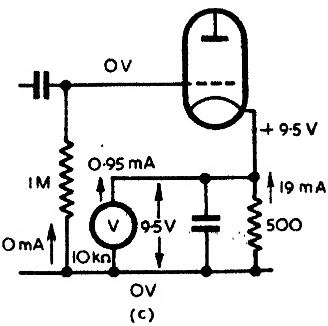


Fig. 3.2—How the measuring instrument may affect the thing measured. (a) shows part of a valve circuit with the conditions of current and voltage marked. If one tries to measure the voltage between cathode and grid with a voltmeter, condition (b) results, in which the quantity to be measured is reduced to one-hundredth of its proper amount. Measuring the voltage across the biasing resistor (c) causes a comparatively small error

Such a method has already been discountenanced on two grounds: it is particularly undesirable to take readings which (as probably in this case) are widely different; and there is the effect of the meter on the input and output circuits. But, in addition, the switching system may introduce enough feedback to affect the performance of the amplifier very considerably. Take great care to reproduce true working conditions, and to exclude disturbing effects due to the measuring apparatus. The approved method of measuring amplification is to connect in series with the amplifier an attenuator which can be adjusted to counterbalance the gain, so that the levels at the input and output of the whole are the same. The attenuation is then a measure of the gain, and no signal-level measurements are necessary.

These examples are given so that when individual methods are discussed later it may be quite clear without lengthy explanation why certain methods are recommended rather than others.

3.14. REPRODUCING WORKING CONDITIONS

The advice to reproduce, as far as possible, natural working conditions in an experiment, and again to study one fact at a time, may tend to conflict. The natural working conditions of a broadcast receiver are that it is installed at a certain place, connected to a certain aerial and earth, turned on, and tuned to certain stations. During that process many influences are at work: the power, locality, carrier frequency, modulation percentage and frequency of the sender; the distance and nature of ground between sender and receiver; the time of day and year; the nature of the receiving aerial and its situation; and the acoustics of the listening-room—to name a few. To draw reliable conclusions about the results, it is necessary to simplify the problem. For instance, the sender may be replaced by a standard signal generator, with everything under control and measurable. The outside aerial may be replaced by a dummy aerial smaller than a matchbox. The loudspeaker may be replaced by an artificial load and output meter. Then some definite figures of receiver performance can be obtained. But they should not blind one to the artificiality of the test, and lead to unwarrantable conclusions.

Lastly, it is necessary to keep a sense of proportion. This is lacking if one goes to great lengths to reduce some relatively minute error when another factor is subject to a much larger one.

CHAPTER 4

Sources of Power and Signals

4.1. " SIGNALS "

In the earliest systems of electrical communication the messages were conveyed by means of make-and-break of current according to a prearranged code: in one word, signals. The name stills persists, although its original meaning has had to be strained severely to make it cover such things as broadcast programmes. But some concise term is indispensable for distinguishing communication currents and voltages from those used for heating valve cathodes, providing anode current, etc. Since an unvarying current conveys no information (to use another word whose original sense has been strained), the essence of signals is variation; and the frequency of variation used in practice covers a very wide range. For a long time there was a broad distinction between audio and radio frequency, the division being at about 20 kc/s; but the range of video frequency used in television and radar covers much of both; so the distinction is not as useful as it was. In the laboratory, controllable sources of a.c. of various frequencies—signal sources—are most necessary items of equipment.

4.2. MAINS POWER

Sources of power for feeding valve apparatus and for incidental purposes are mainly d.c., although the raw a.c. from the mains is used wherever practicable. Elsewhere, rectifiers and filter circuits are needed to convert to d.c. Generally it is more convenient to build a power unit into each individual piece of apparatus that requires it than to try to run it from some central source. The requirements are so varied that it is impossible to design a power unit to anticipate them all. If it is larger than is needed not only is there waste, but also, in order to reduce the voltage without bad regulation (i.e., change of voltage for varying currents drawn), expensive stabilizing devices may be necessary. And if common power supplies are used for a number of instruments, there is a probability of short-circuits and undesirable coupling. There are other obvious advantages in providing tailor-made power units.

Nevertheless it is very convenient to have in addition one or two " off the peg " for general purposes; one giving up to 120 mA d.c. at 350 V and several amps of a.c. at 6·3 V (preferably from more than one winding) is very useful. Details of design need not be given here because they follow ordinary receiver practice, and can be varied to suit individual needs. Several transformers supplying a.c. at 6·3 V, with tappings at centre, 4 V and 5 V, are useful for heating valves connected in experimental circuits, and sometimes come in handy as a source of 50-c/s signal voltage.

A larger transformer giving several hundred watts at any voltage up to 250 or 300 is a useful possession. If varied by a tapping switch (with intermediate dead studs to prevent short-circuits) it may be necessary sometimes to use a heavy-current sliding potential divider across it for fine control, but the Berco " Regavolt " and the General

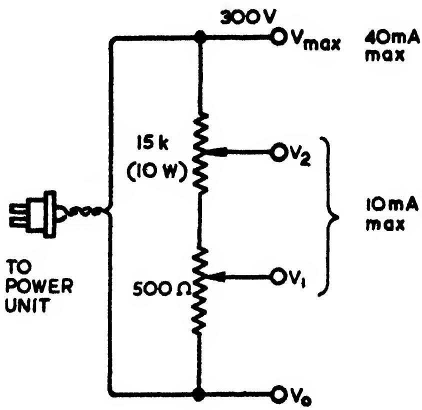


Fig. 4.1—Potential divider for extending the usefulness of a power unit—in this example, 300 V

Radio " Variac " combine both functions with greater efficiency, and give a practically continuous variation of voltage. Less convenient, but cheaper, is a switched transformer giving an output in steps of 0·5 or 1 V, from 1 to 255 or 511 V, for which constructional details are given by H. E. Styles (*W.W.*, June 1958).

The ability to control output voltage over a wide range, yet very precisely, is a most valuable feature in a laboratory power unit. It can conveniently be added to an ordinary uncontrolled power unit by making up some kind of variable potential divider to plug in to it between its output terminals. No very simple arrangement is suitable for supplying a range of current from zero up to the full output of the power unit at any fraction of full voltage; a useful compromise is a fairly heavy potential divider, itself drawing perhaps a quarter to a half of the total output power, and rated to stand rather more than that, restricted to supplying a similar fraction of its own consumption. Fig. 4.1 shows a typical example for a power unit giving a maximum of 60 mA at 250–300 V. With V_1 joined to V_0, the V_1 control can be used as a fine control of V_2. A much better control is obtained by using cathode-follower rectifiers in the power unit, as explained by A. H. B. Walker (*W.W.*, Sept. 1952).

4.3. VOLTAGE STABILIZERS: REFERENCE DEVICES

For many purposes it is desirable, or even essential, for the output voltage of the power unit to be stabilized (or regulated) against variations of input voltage and output current. Some of the means used

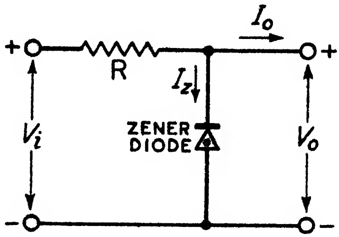


Fig. 4.2—The simplest type of voltage stabilizer: a voltage reference device and a resistor

for doing so can be elaborated in such a way as to provide control of the output voltage over a considerable range. Units of this kind are extraordinarily useful and are commercially available in great variety.

Almost the only feature common to all of them is a device across which the voltage remains nearly constant in spite of variations in current, temperature, etc. A constant type of battery could of course be used (Sec. 6.12), but more usual is a gas-filled diode called a voltage reference or stabilizer tube. A typical tube " strikes " at about 115 V, which immediately drops to about 85 V. Its incremental resistance is about 300 Ω, which means that over a current range of 1 to 10 mA the voltage changes by only about 2·7 V. This change tends to take place in small jumps rather than smoothly. The temperature coefficient is about −4mV/°C.

Unfortunately the choice of voltages is very limited, especially if a high standard of constancy is required, and the lowest is about 55 V. This limitation is absent from certain diodes known as Zener diodes, which are operated in what with other diodes would be regarded as the wrong direction, beyond the voltage breakdown point. Another advantage is that these diodes " strike " at normal conducting voltage; there is no switching-on surge. Zener diodes are obtainable for almost any voltage from 2 V upwards, but below about 5 V the transition from not-conducting to conducting is not very sharp. They are especially useful for about 5–6 V, at which the temperature coefficient is least. It is negative below that, and positive above; of the same order as in gas tubes. Incremental resistances are of the order of 10 Ω, and types are available with working currents of 500 mA or more.

4.4. SIMPLE STABILIZERS

Any of these devices can be made into a simple voltage stabilizer by means of a suitable resistor, as in Fig. 4.2. If the input voltage

V_i increases, more current passes through the diode, but nearly all the resulting voltage increase occurs across R. And if more current is drawn from the output terminals, that much less passes through the diode, provided of course that there was enough originally.

Fig. 4.3 shows an example of a Zener characteristic curve, which would be suitable for supplying current at a nominal 6·3 V. The choice of R depends on the extent to which the input voltage V_i and the output current I_o are liable to vary. Supposing V_i may vary between 8 and 10, and 40 mA is the maximum allowable Zener current (I_z), and the minimum I_o is zero, what is the maximum I_o that can be drawn without ever going beyond the limits of regulation? The conditions giving the highest I_z would be maximum V_i and zero I_o, so draw a line from 10 on the voltage scale to 40 mA on the Zener curve. Its slope turns out to represent 85 Ω. At the minimum V_i (8 V) a line with the same slope cuts the curve at 20 mA. The minimum I_z on the "working straight" is about 2 mA. So (neglecting the variation in voltage across R) the maximum I_o that can be drawn,

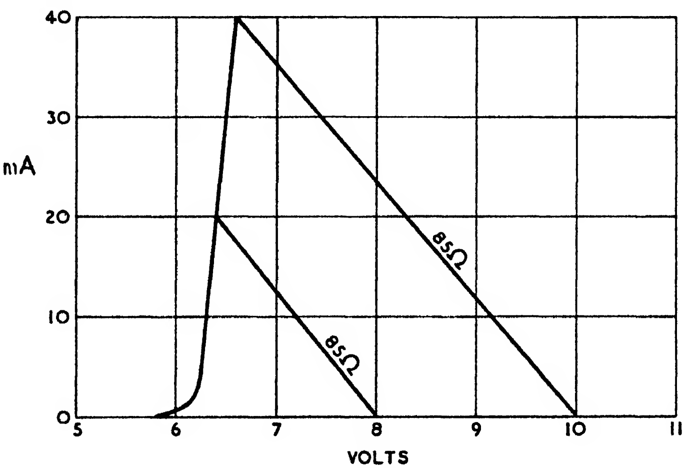


Fig. 4.3—Typical Zener diode characteristic curve, with series resistance lines drawn to show performance with supply voltages of 8 and 10

under this worst V_i condition, is 18 mA. The rise of V_o for a V_i rise from 8 V to 10 V is from 6·4 V to 6·57 V, so the ratio of the percentage rises (which is a measure of the voltage stabilization) is 9·4. The Zener slope shows an incremental resistance of about 8 Ω, so the effective source resistance (8Ω in parallel with 85 Ω) is 7·3 Ω. A larger output current could be regulated if it could be guaranteed not to fall below a certain minimum; R would in that case have to be correspondingly lower.

For more information on Zener diodes and their applications, see

J. M. Waddell and D. R. Coleman, *W.W.*, Jan. 1960, and J. A. Chandler, *E.E.*, Feb. 1960 and July 1960, p. 449.

Calculations for gas tubes are complicated by the fact that the striking voltage is higher than the running voltage; design is discussed in detail by J. W. Hughes (*W.E.*, Aug. 1947).

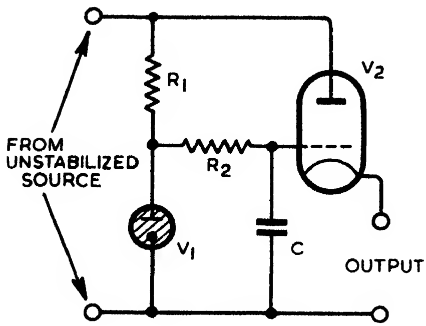


Fig. 4.4 (left)—Voltage reference tube and cathode-follower voltage stabilizer

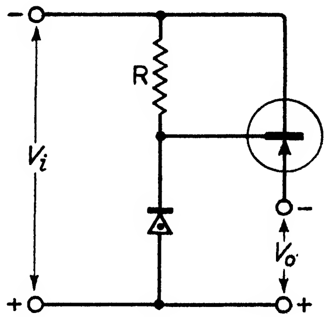


Fig. 4.5 (right)—Semi-conductor analogue of Fig. 4.4, in which a transistor takes the place of the valve, and a Zener diode replaces the gas tube. With a pnp *transistor, as shown, the polarity is reversed; with an* npn *transistor and reversed diode it would be the same as with a valve*

If the regulated current obtainable by use of the reference device as a direct shunt is insufficient, it is possible to extend it by a valve connected as a cathode follower (Fig. 4·4) or a transistor connected as an emitter follower (Fig. 4.5). For the higher voltages, valve and gas tube naturally work together, and the striking voltage " spike " is suppressed by the warming-up period of the valve. Another advantage is that tube " noise " (rapid fluctuations), which for some applications may be appreciable, can be reduced by R_2C. The values are not critical; a time constant of 0·1 sec is about right. The output resistance of a cathode follower is approximately $1/g_m$. This type of stabilizer circuit is discussed by A. P. Willmore (*E.E.*, Sept. 1950). Design of the analogous diode and transistor circuit is illustrated by S. Welldon (*W.W.*, Aug. 1958).

4.5. STABILIZERS WITH AMPLIFIED CONTROL

The stabilization ratio in the foregoing arrangements is usually not more than about 10, but it can be pushed up into the hundreds or even thousands, and the output resistance reduced to less than 1 Ω, by amplifying differences between the output voltage and the reference voltage and applying them to the grid of the cathode follower. The design of this type has been discussed fairly fully by the author (*W.W.*, Oct. to Dec. inclusive, 1948); the following is no more than an outline.

In Fig. 4.6, where for the time being R_4 should be regarded as open-circuited and R_5 short-circuited, V_1 corresponds to the cathode follower, through which the whole of the load current passes. This valve serves two purposes: it takes across itself the variations of input voltage (due to load and mains variations), thereby preventing them from being passed on to the output; and it enables the output voltage to be varied. As the difference between maximum and minimum output voltage, plus a cushioning voltage to deal with the maximum variations, plus the

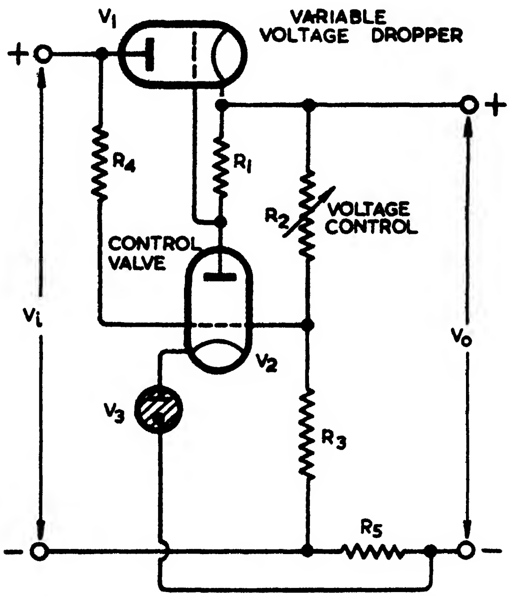


Fig. 4.6—Outline circuit of a power-supply voltage regulator for obtaining a stabilized voltage adjustable over a wide range. R_4 and R_5 are supplementary devices for improving the stabilization

unavoidable drop across the valve when carrying full load current with negative grid, may be considerable, V_1 must have an adequate power dissipation rating.

The voltage drop across V_1 is controlled by negative grid bias derived from R_1, in the anode circuit of an amplifying valve V_2. Unwanted variations in output voltage (or at least as large a fraction of them as possible) are passed to the grid of V_2, and appear on an amplified scale across R_1, in such a sense as to tend to suppress the variations. Output voltage can be controlled by adjusting R_2.

The reference tube V_3 raises the cathode potential of V_2 by a fixed voltage, so that variations in grid voltage are wholly effective for control purposes, and are not largely offset by cathode-follower action.

It is obvious that the arrangement just described is incapable of suppressing variations in output voltage completely, seeing that its correcting action depends on the existence of such variations; there is also some loss of efficiency because of the potential-reducing effect of R_2R_3. Two of many possible improvements are shown.* Variations of *input voltage* are fed via a high resistance R_4 to the grid V_2, thereby aiding the suppressing action in V_1. Variations of *output current*, which tend to affect the voltage, are passed through the low resistance R_5, and by affecting the cathode potential of V_2 have a similar correcting effect. R_4 and R_5 are adjusted on test to give optimum correction over the range of output voltage. With the circuit shown, that range can hardly begin at a lower voltage than about 125; but its top limit is fixed only by availability of suitable valves and power supply voltage.

Input compensation introduced at the suppressor grid of the amplifier has been shown to give good results (Chatterjee, Ramanan and Rao, *E.E.*, May 1961).

4.6 DESIGN PROCEDURE

The procedure for determining suitable values of components and types of valves is explained in detail in the articles referred to. Fig. 4.7 is an example of the type of diagram devised by the author for dealing with the series valve—actually in this case a pair of triode-connected EL37 valves. Voltage is plotted against current passing through these (i.e., output or load current, I_0). Suppose the intended maximum output voltage (V_0) is 400 V. This, at constant output current, is represented by the horizontal line AB. Maximum I_0, say 100 mA, is represented by point B. The I_a/V_a curves for the EL37 show that at 50 mA each, and minimum currentless V_g (say $-1\frac{1}{2}$ V), their V_a is about 70. This is represented by BC. From C is drawn a sloping line CD which is the regulation curve of the unstabilized power unit at the minimum mains voltage, say 90% of normal. Point E can then be plotted, at 100/90 times the voltage of D, to represent the unstabilized output at normal mains voltage; and F at maximum (say 105% of normal). EG and FH are then drawn parallel to DC.

The vertical distance between the appropriate sloping V_i line and the horizontal V_0 line is the voltage V_a across the valve. One can transfer the I_a/V_a curves from the valve data sheet to this diagram, hanging them as it were from the V_i lines, and so find the required V_g at any output current; but it would be rather confusing to do so on one diagram for all three mains conditions. The dotted line DB is a sample ($V_g = -1\frac{1}{2}$ at 10% low mains) and corresponds to full I_0 (point B). As I_0 is reduced towards A, contact is made with valve curves for increasingly negative V_g. What one has to find is the extreme range of V_g to be provided. The minimum, which occurs at minimum V_i (C) and maximum V_0 (B) has already been decided: it is $-1\frac{1}{2}$ V. The maximum

* Due to Lindenhovius and Rinia. *Philips Technical Review*, Feb. 1941.

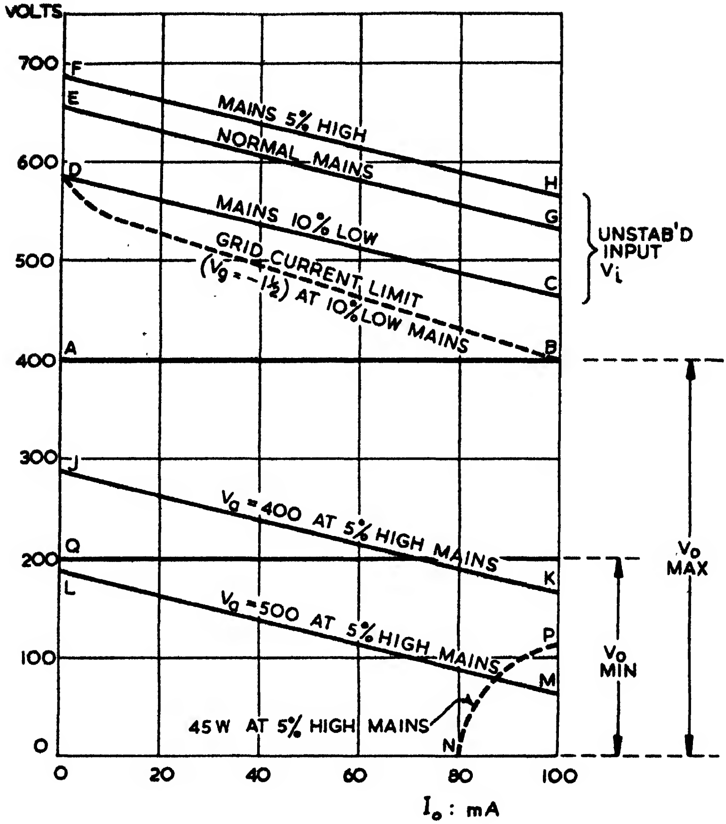


Fig. 4.7—Example of design chart for series-control voltage stabilizer

occurs at maximum V_i (F) and minimum V_0, which has yet to be chosen. If the maximum allowable V_a is taken as 400 V, then this is represented by JK, drawn 400 V below the highest V_i line, FH. This would restrict the minimum V_0 (at $I_0 = 0$) to 290 V; but if 500 V were allowed (LM) a minimum V_0 of 200 would be within the boundary. Another possible restriction is allowable anode (and screen) dissipation. With the EL37 it is 28 W each, but it would be prudent to assume no more than 45 W per pair, to allow for unequal load sharing. The V_a corresponding to this limit at several values of I_0 is calculated, and set off below FH, giving the curve NP. This is well below any practical V_0 level. Finally, if 200 V is to be the minimum V_0, the maximum required V_g, corresponding to point Q, is found from the valve data for $I_a = 0$ and $V_a = 490$ (QF), and turns out to be about -70 V.

Next, the valve curves for V_2 are studied to find the input required to give this range of $-1\frac{1}{2}$ V to -70 V. Connected as in Fig. 4.6, the system has a very low anode current for V_2, and hence low amplification, at the 1½-V end. A great improvement is obtained by feeding V_2 from a point more positive than the cathode of V_1, as shown in Fig. 4.8. The EF80 is very suitable for V_1 and 85A2 for V_3. The screen of V_2 can be fed from a potential divider as in Fig. 4.8, and by a suitable choice of values this can be made to serve as an alternative to R_4 for input-voltage compensation. Hum is reduced by connecting about 0·5 μF across R_2. If valves in parallel are used for V_1, care must be taken that they share the load as equally as possible, or the one taking more than its share is likely to fail and cause the others to go too.

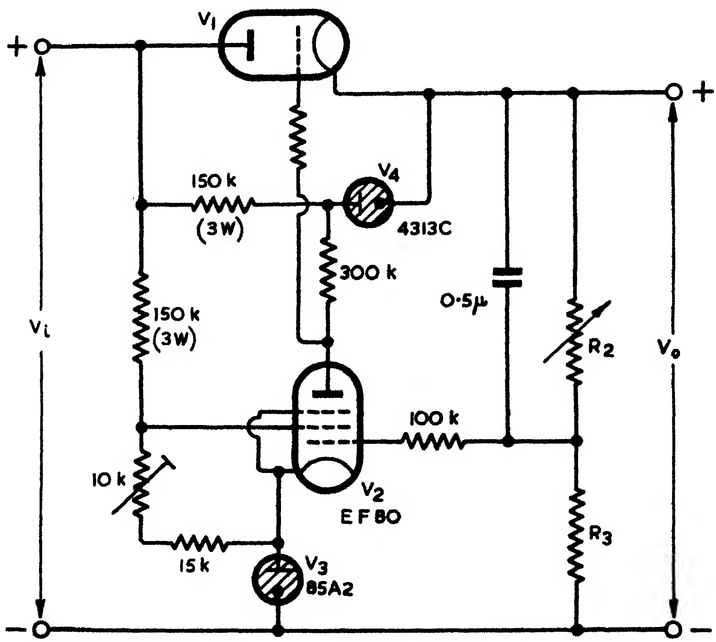


Fig. 4.8—Modified circuit for obtaining a high degree of stabilization over a wide range of voltage. The V_1 grid circuit resistance is not critical; a typical value is 1kΩ

Resistors in series with the grids of V_1 and V_2 are advisable. It is normally necessary to provide independent heater windings on the supply transformer for V_1 and V_2. Precautions must be taken to prevent V_2 from picking up hum from the power supply or elsewhere. Its input is very sensitive to small voltage changes, and it is essential for R_2 and R_3 to be very stable; R_2 should certainly be a wire-wound variable covering the required range, in series with a fixed resistor.

When properly made and adjusted, such a voltage regulator looked at from the output terminals is an almost zero impedance in series with a z.f. voltage. With careful design there is no reason why a unit on these lines should not have an equivalent internal resistance of $< 1\Omega$ and hum < 1 mV. Rapid fluctuations in mains voltage can be reduced by a factor of the order of 3,000, but if a voltage change is maintained there is a relatively large slow drift due to varying heater temperature. For more effective protection against large changes it is necessary to stabilize heater voltage too (Sec. 4.11).

4.7 MODIFICATIONS

Some design data on the type of stabilizer just described are given by F. A. Benson (*E.E.*, March 1952) and (with L. J. Bental in *E.E.*, Sept. 1956) a review in particular of methods (including the present author's) for obtaining a wide range of voltage variability. A brief review by D. J. Collins and J. E. Smith, including certain improvements chiefly for operating the voltage reference tube more favourably, appears in the April 1959 issue. A design for fulfilling extremely stringent requirements where battery power had previously been considered indispensable was given by Dr. D. L. Johnston (*W.W.*, Sept. and Oct. 1947).

Very high voltage amplification, of the order of 1,000 times, can be obtained by substituting for the pentode (V_2 in Fig. 4.8) a double-triode cascode stage using a final anode coupling resistance of 2·2 MΩ as advocated by V. H. Attree (*E.E.*, April 1955). He claimed an output impedance as low as 0·2 Ω and hum less than 0·5 mV, without any compensating devices. But that such a high value of coupling resistance may restrict the operation of V_1, and perhaps fall outside the maker's rating, was pointed out by P. J. Franklin, P. D. Neville and J. L. Thomas (*M.T.C.*, Nov. 1958), in a paper which gives detailed data on the parallel running of series valves, with particular reference to the 6080 double triode.

Maintenance of constant voltage implies a risk of current overload if the output terminals are connected through too low a resistance or too large a capacitance. C. R. Cosens has shown that a safety device can be added at low cost, as in Fig. 4.9. When the load current exceeds a certain limit, decided chiefly by R_k, the diode conducts and prevents any further considerable increase in current even if the output is short-circuited. A typical value of R_k is 220 Ω, which admittedly adds to the r_a of V_1, but the stabilization is only slightly reduced. More recently a similar arrangement has been described by D. P. C. Thackeray (*E.E.*, Nov. 1958).

In the same journal, June 1945, p. 559, E.M.I. Laboratories showed a modified circuit in which a negative as well as a positive stabilized supply is obtained without using any more valves, but at some sacrifice of performance and flexibility.

The manually controlled cathode-follower rectifier system by A. H. B. Walker, referred to at the end of Sec. 4.2, can be combined with automatic regulation to give an economical unit of lower performance. This type has been discussed by B. J. Perry (*E.E.*, Dec. 1956). The design presented briefly by D. W. W. Rogers (*E. & R. E.*,

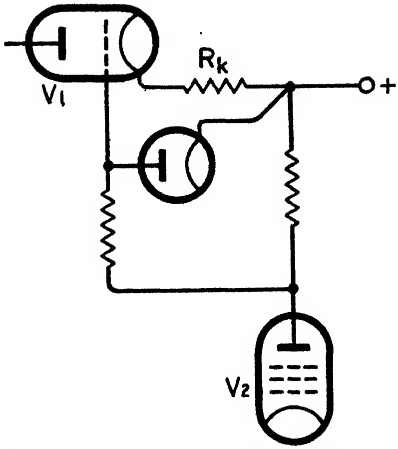


Fig. 4.9—Automatic current-overload preventer for voltage stabilizer, comprising the diode and R_k

Sept. 1957), in an article devoted mainly to minimization of noise in the more conventional type using a separate rectifier, is however intended for a high-frequency mains supply.

Where the d.c. to be provided exceeds, say, 0·5 A, thyratrons can economically be used as the combined rectifiers and control valves, up to as much as 70 A. The design of such units, with and without control amplification, was dealt with by B. G. Higdon and M. E. Bond (*E.E.*, Aug. 1959).

In the type of stabilizer considered so far, a valve in series with the load absorbs surplus voltage. The shunt type, in which a valve in parallel with the load by-passes surplus current, may offer a simpler solution in some situations. Its design was treated by J. McG. Sowerby (*W.W.*, June 1948).

4.8. STABILIZED TAPPINGS

One often wants to draw off small currents at a lower voltage than the main stabilized supply; for screen-grid voltages, say. The ordinary high-resistance (0·1 MΩ order) potential divider effectively unstabilizes the voltage it provides, and a low-resistance one would waste precious stabilized current. The answer is the directly controlled cathode follower already seen (Fig. 4.4) but with its grid connected to a high-resistance potential divider across the stabilized supply. Incidentally, if this divider's resistance is as constant as it should be, it can be made to serve also as the series resistance of a voltmeter for the main supply. Using a high-slope valve, one achieves an output resistance of only one hundred ohms or even less, which is usually low enough for small currents fed to individual circuit branches. With a variable potential divider the voltage of the tapping can be adjusted over nearly the whole range of the supply to which it is connected, but care must be taken not to exceed the ratings of the valve.

The design of these cathode-follower tappings has been discussed more fully by the author (*W.W.*, Jan. 1949).

4.9. LOW-VOLTAGE STABILIZERS

Although it is possible to design valve-operated stabilizers so that the voltage can be varied right down to zero, in general they are unsuitable for the low voltages needed by transistors, for example. This is the more so because the current requirements are likely to be correspondingly heavier.

For many applications the direct use of a Zener diode (Sec. 4.4) is satisfactory, either with one " stage " as in Fig. 4.2 or with two or more. But where the requirements include variable voltage or high constancy or currents of the ampere order, one looks for transistor-operated units analogous to the valve units already described. The simple emitter-follower type (also Sec. 4.4) goes a very little way to meeting the more exacting needs.

We have seen that simple valve circuits such as Fig. 4.6 are capable of a very good performance, and might suppose that the valves could quite easily be replaced by transistors. But nowhere do the differences between valves and transistors show up more clearly.

When the series valve is used to reduce the output voltage from a fixed input, it must itself dissipate the surplus power. Compared with a valve, a transistor is not a good dissipator of power. It can handle heavier currents and moderately high voltages, but not both at once.

Then a transistor differs from a valve in that its base voltage has to be intermediate between collector and emitter, so that if one were substituted for V_2 its output would be of the wrong polarity. If, to put this right, it is taken from the unstabilized side of the series transistor, its amplification is largely counteracted and the stabilization suffers accordingly. So either more stages have to be added or an extra stabilized supply arranged for the transistor.

The low input resistance of transistors and their variability with temperature are other differences that the designer has to take into account.

The net result is that even for a performance that would be ordinary in a valve unit, transistor units tend to be complex and expensive. One can evade the worst difficulties by having a fixed output voltage, but for that one might almost as well use batteries and save a lot of expense and the inconvenience of a mains connection. It must be remembered that while a power supply impedance of 1 Ω is tolerable in most valve circuits, at the low voltages and sometimes large peak currents of transistor circuits a much lower value is needed.

The circuit of the Cossor Model 1329 power unit, Fig. 4.10, is about as closely analogous to its valve counterpart as can be, and at the same time illustrates the differences. It provides 0 to 2 A at a voltage variable from 5 to 10. The stabilization ratio is given as

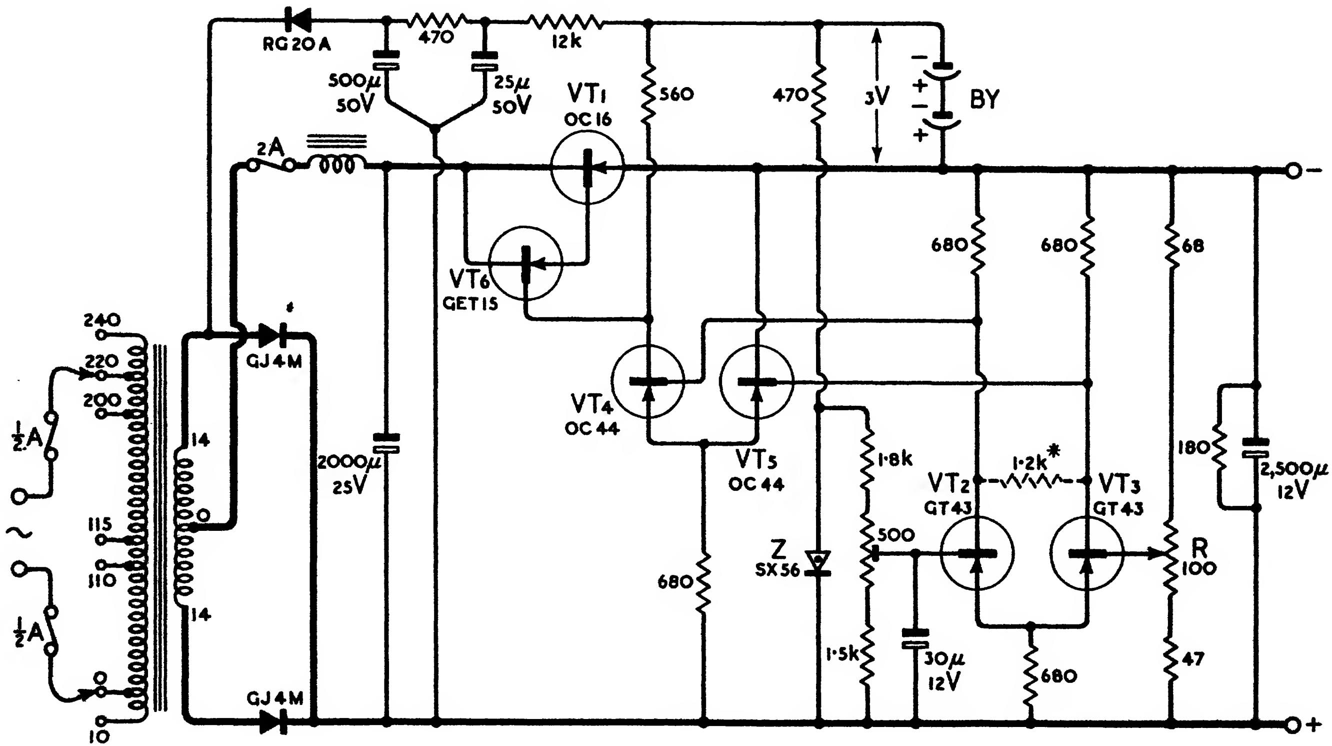


Fig. 4.10—Circuit diagram of Cossor 1329 stabilized power unit. Note the auxiliary supply, stabilized by BY, to provide a more negative potential than that of the main stabilized output. The item marked is a compensating resistor, to be used if required*

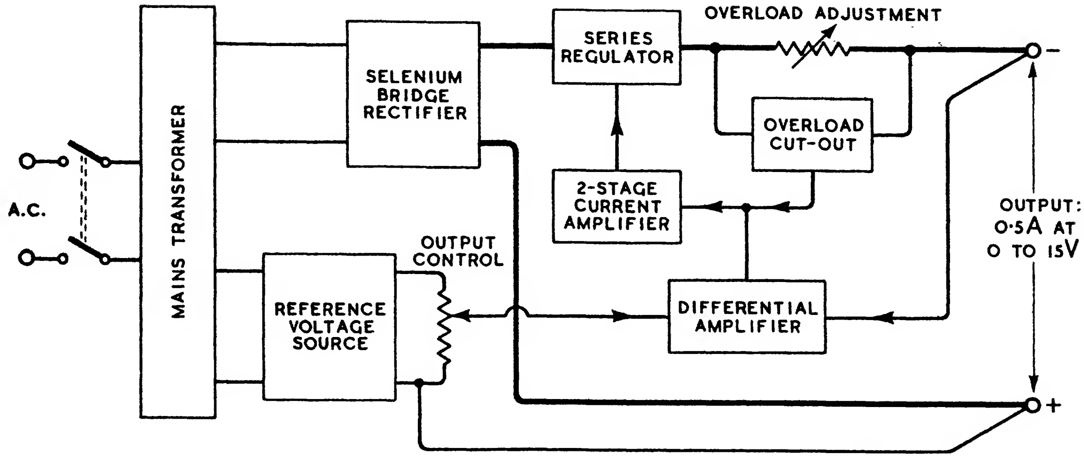


Fig. 4.11—Block diagram of Advance PP5 transistor stabilized power unit

200 : 1, the ripple as not greater than 1 mV peak-to-peak, the temperature coefficient not greater than 1 mV/°C, and the output impedance less than 0·01 Ω. Transistor VT_1 corresponds to series valve V_1, and VT_{2-6} correspond to amplifier valve V_2. There are two stages, each consisting of coupled-emitter " long tailed pairs " for stability (especially against temperature variations), and one emitter follower (VT_6) to provide a low output resistance to match the input of VT_1. The first stage compares a proportion of the output voltage, controllable by R, with a proportion of the reference voltage across a Zener diode, Z. This diode is a 5·6 V type for minimum temperature coefficient, and, as 5·6 exceeds the minimum output voltage, a more negative point has to be provided by means of two cell-type stabilizers, BY. These also provide the bias for VT_6 and VT_1.

The Advance Type PP5 unit is similar in its essentials (Fig. 4.11); its output is 0·5 A at a voltage variable from 0 to 15. A valuable feature is an overload cut-out, adjustable to disconnect the supply in 0·2 millisecond at anything from one-tenth to full load current.

For variable-voltage heavy-current units it is necessary to adopt some expedient to avoid having to use an excessive number of expensive power transistors. One method, seen in the higher-powered Advance units, is to vary the input voltage to the rectifier in steps. Another, used in Solartron units and explained by D. J. Collins and J. R. Pearce (*E.E.*, Feb. 1960), is to shunt the series transistors with a resistor. In certain circumstances the number of such transistors can be reduced (for example) from ten to four.

Yet another method is to abolish the series transistor(s) altogether, using instead a transductor—an inductor, the impedance of which can be varied by magnetic saturation of the core by current through a control winding—in series with the supply transformer. An example of this technique is the Claude Lyons PST series of units, with outputs ranging from 8 A 12 V to 2 A 48 V. The control current in the transductor is automatically adjusted by a transistorized amplifier monitoring the output voltage. A valuable feature of this kind of unit is its inherent limiting action which prevents full load current being much exceeded even through a dead short.

References on design:

" The Design of Direct Voltage and Current Stabilizers using Semiconductor Devices ", by D. G. Wenham. *Proc. I.E.E.*, Vol. 106, Part B, Supp. 18, May 1959.

" Battery-powered Marine Radar ", by L. H. Dawson. *W.W.*, Aug. 1960. (For a stabilizer producing 4 to 8A at a constant 19V, from a supply varying from 20·5 to 32V.)

" A Circuit for the Protection of a Stabilized Transistor Power Supply ", by H. Kemhadjian and A. F. Newell. *E.E.*, April 1960.

4.10. CURRENT STABILIZATION

Although d.c. power supplies are usually required to approximate to constant-voltage sources, there are occasionally calls for constant-current sources. Here again, if the requirements are not too stringent, the versatile cathode follower comes to our aid, basically as in Fig. 4.12 (a). The load can alternatively be placed in the negative line, but there is then greater difficulty in providing the stabilized grid voltage, which in the arrangement shown can easily be done by means of a gas or Zener diode, fed through a suitable resistance from $+V_b$.

By cathode-follower action the cathode potential is held nearly constant at a few volts more positive than the grid, so the anode current is determined almost entirely by the value of R and only slightly by the load resistance. The device is, in fact, equivalent to a resistance of $R_a + (\mu + 1)R$ in series with $V_b + \mu V_g$ volts, where R_a is the d.c. resistance of the valve. For the sake of high μ a pentode is obviously

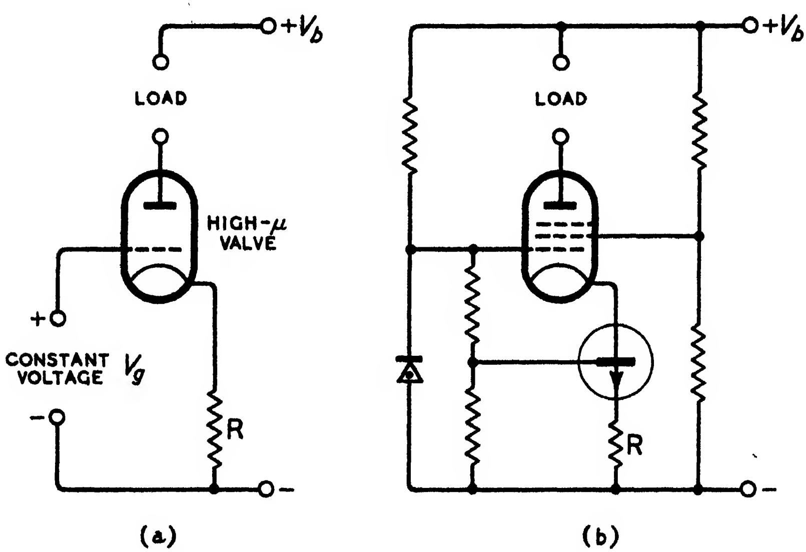


Fig. 4.12—Current stabilization can be obtained by connecting a cathode follower as shown (a) so that the load sees it as a high incremental resistance. To multiply the effect of R without a correspondingly large voltage drop, an emitter follower can be used as at (b). Here a Zener diode is shown keeping the grid voltage constant

attractive, and with an anode-connected load its screen-grid could be maintained at a nearly constant voltage by the same means as the control grid.

While an ordinary resistor can be used for R, especially if the current is to be adjustable, better stabilization for a given voltage drop can be obtained by using a pentode or (for lower values of V_g) an *npn* transistor, itself connected as a current stabilizer (Fig. 4.12 (b)). With $R = 500\Omega$, $I_a = 10$ mA, and the cathode only about 8 V positive, the effective resistance in series with the load can be greater than 50 MΩ. As in all stabilizer circuits employing transistors, care is needed to ensure that no limit ratings are exceeded under any possible working conditions.

Other current stabilizer circuits are shown by J. H. McGuire (*E.E.*, Dec. 1955), by L. N. Clarke (*E.E.*, May 1956, p. 223), and B. G. L. Braun (*E.E.*, Aug. 1956, p. 360).

4.11. STABILIZATION OF A.C.

The need to stabilize the heater supplies of valves in stabilizer circuits which are required to give very precise compensation has already been

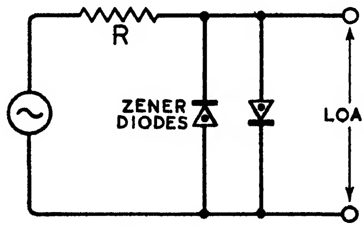


Fig. 4.13—Simple voltage stabilizer for a.c. Because R must be large, it is wasteful of power

mentioned. But of course there is other work that is liable to be upset by variations in mains voltages, especially when the public supply system is overloaded. The best thing no doubt is to stabilize the entire 50-c/s supply to the laboratory; and firms such as Zenith and the British Electric Resistance Co. have automatic servo-operated regulators rated up to 20 kVA or so.

For the range 30 to 1,200 VA maximum, there is the Westinghouse " Stabilistor ", described by A. H. B. Walker (*W.W.*, Nov. 1944), and very fully by the makers in their Rectifier Data Sheet No. 901. Although the action depends on saturated iron-core inductors, the distortion usually associated with this class of stabilizer is largely avoided. Two types are available: A, which stabilizes very effectively against mains voltage and load variations, but not frequency variations; and B, which is proof against frequency variations but must be run at constant resistive load.

The low-distortion constant-voltage transformers by Blackburn Electronics Ltd., with a stabilization ratio of 200 : 1, can be used in front of electronic instruments with their own unstabilized supplies, as an alternative to d.c. stabilization.

The matter of waveform distortion is important for many purposes, among which valve heater supplies might be overlooked. It has been pointed out that they need constant *power*, which is unlikely to be obtained if the peak voltage (or current) is kept constant while the waveform is allowed to vary.

The very simple arrangement of Fig. 4.13 can be made to give nearly constant power output if *R* is high enough for a nearly square output waveform to result, but of course the power efficiency is then very low.

G. N. Patchett has devised several types of a.c. voltage stabilizer; one, in which a thermistor is used to control a valve-loaded transformer in series with the supply, keeps the output voltage within the extremely narrow limits of 0·01%, with harmonic distortion less than 1%. The design is fully described by him (*E.E.*, Sept. to Dec. inclusive, 1950).

For less precise control by simpler means, F. A. Benson favours a temperature-limited diode as the controlling element (*E.E.*, June 1956).

For stabilizing the valve heaters in a low-power instrument voltage stabilizer, C. Morton has shown a system in which the stabilized d.c. itself, 100 mA, is used to feed the heater in series (*E.E.*, Feb. 1952).

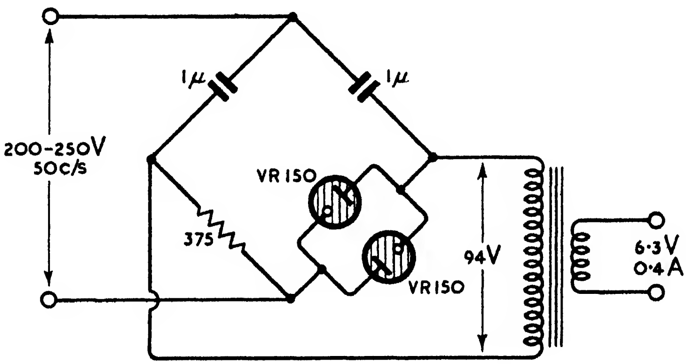


Fig. 4.14—Cherry & Wild bridge stabilizer for a.c., with circuit values suitable for feeding a valve heater

A very simple and cheap arrangement giving several watts of constant-voltage a.c. is described by L. B. Cherry and R. F. Wild (*Proc. I.R.E.*, April 1945). Circuit values used by the present author are shown in Fig. 4.14. The variation in output voltage for an input voltage variation of ±10% is given by Cherry and Wild as ±0·35%, and the power efficiency 30–50%.

A.C. stabilization in general is reviewed by O. E. Dzierzynski (*W.W.*, Oct. 1957).

G. N. Patchett reviews all kinds of a.c. and d.c. voltage stabilizers in *Automatic Voltage Rectifiers and Stabilizers* (Pitman, 1958); and in *Voltage Stabilized Supplies* (Macdonald, 1957) F. A. Benson includes no fewer than 1,014 references to literature on the subject.

4.12. BATTERIES

Just when mains power units had almost completely displaced batteries from the laboratory, the development of very low consumption valves and (still more) transistors began a revival of batteries. As the last few Sections have shown, stabilized power units are not at their best in the low-voltage high-current sizes. While batteries need periodical maintenance or replacement, and their voltage cannot be varied continuously, they are cheaper, simpler, smaller, have less stray capacitance and are free from ripple and from bondage to the mains.

For low internal resistance, and current running into amps, the lead/acid accumulator is the usual choice. It will not stand neglect, and should have regular and frequent use and maintenance (but not close to apparatus liable to be affected by acid creepage or fumes). Its normal e.m.f. (2 V per cell) is convenient for mental arithmetic but perhaps rather large as a step of variability.

Although the Venner silver/alkali accumulator* gives a less constant voltage and is much more expensive, it has a fraction of the weight and can be allowed to remain discharged indefinitely. It gives 1·5 V nominal.

Of primary batteries, the so-called dry Leclanche type is cheap and universally obtainable in many sizes and shapes. Its voltage drops considerably from the nominal 1·4 V if discharged continuously at even a low rate; and it deteriorates with storage, though perhaps less so than is sometimes supposed, for in the larger sizes at least and at reasonable temperatures it is capable of occasional use over many years. But on no account must it be left inside apparatus in a discharged condition, as it is liable to burst and make a very corrosive mess.

The mercury/alkali cell has much in its favour for laboratory purposes, especially for small portable equipment, on account of its long storage life and constancy of voltage. Indeed, it has been used successfully in place of standard cells, with great saving of cost. At 70°C, a maker's† curves show that on storage the initial e.m.f., 1·357 V, drops only $\frac{1}{4}\%$ in 11 months. At 1 mA continuous discharge, a cell rated at 1,000 mA-hr dropped to 1·32 V during the first 60 hours; then remained almost constant at that for 900 hours. Mercury cells are obtainable in a wide variety of sizes and characteristics.

Modern battery developments were reviewed in *W.W.*, Dec. 1958.

4.13. D.C. VOLTAGE RAISERS

It is convenient, and sometimes essential, to use instruments that need no mains connection. Where only transistors have to be fed, a battery with very few cells can be used. Valves need many cells, which must be very small and thereby short-lived; and cathode-ray tubes need a generally impracticable number of cells. Portable instruments such as oscilloscopes are, however, now obtainable because the substitution of transistors for valves has greatly reduced both voltage and power required, and at the same time transistors enable the battery voltage to be stepped up for a c.r. tube or other high-voltage load.

The principle is to back-couple a transistor by means of a transformer winding, causing it to generate a nearly square wave at a frequency of, say, 1,000 c/s. Another winding on the transformer

* Venner Accumulators Ltd., Kingston By-pass, New Malden, Surrey.
† Mallory Batteries Ltd., Rainham Road South, Dagenham, Essex.

steps up the voltage to the desired amount, which is rectified and smoothed in the usual way. Fig. 4.15 (a) shows the essentials. Because the transistor works virtually as a switch, maximum current in the "on" position being accompanied by nearly zero voltage drop across it, the efficiency can be high—over 80%.

This type of circuit can be elaborated by incorporating the usual voltage-multiplying device, as in Fig. 4.15(b). Miniature silicon

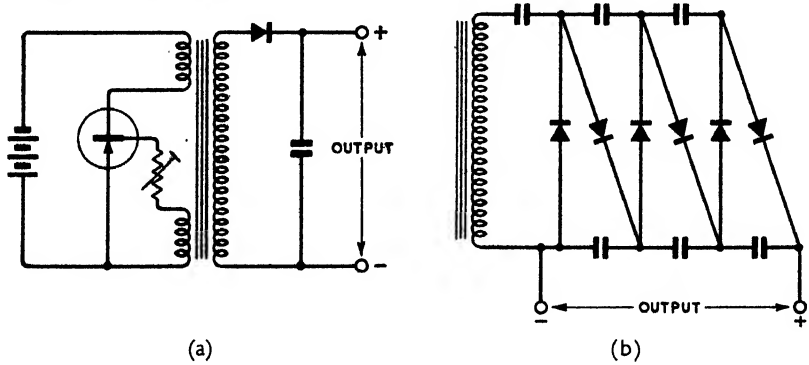


Fig. 4.15—(a) Basic circuit of transistor d.c. voltage raiser, comprising oscillator, step-up transformer, rectifier and smoother. (b) A further step-up from the transformer secondary can be obtained by connecting rectifiers in voltage-multiplier formation

rectifiers are used, as it is necessary for the reverse current to be very small; e.g., 0·2 μA. As much as 10 kV can be obtained from a 1½ V cell.

Stepping up the voltage of d.c. can be combined with voltage stabilization in one unit. All-transistor designs for current of the order of 0·1 A at ±300 V, from a 27·6 V d.c. source, have been given by J. S. Bell and P. G. Wright (*E.E.*, Dec. 1960).

Refs.: "Transistor Power Supplies", by L. H. Light. *W.W.*, Dec. 1955.
"Transistor Inverters and Converters", by M. D. Berlock and H. Jefferson. *W.W.*, Aug., Sept. and Oct. 1960.
Papers by C. G. Avis, C. J. Yarrow and T. Konopinski in *Proc. I.E.E.*, Vol. 106, Part B, Supps. 16 and 18, May 1959.
"The Choice and Design of D.C. Convertors", by J. S. Bell and P. G. Wright. *E.E.*, April 1961.

4.14. SIGNAL SOURCES: THE GRAMOPHONE

For the amateur an inexpensive and practical audio source is the record player he already has. Besides pure tones of fixed frequency he can get fancy test signals—warble tone, gliding tone, etc.—and, of course, any sort of programme as well. Details of test records are given in Sec. 14.38. Such records are of course essential for tests on pick-ups.

Obviously the correctness of the frequency given by a record depends on the motor speed. Assuming that the mains are frequency-controlled a.c., the speed can be accurately adjusted by means of a stroboscope card—or the requisite number of black and white stripes painted on the edge of the turntable—illuminated by a lamp; a neon lamp is preferable as it is extinguished completely twice each cycle of supply voltage, whereas the flicker of a metal-filament lamp is very slight. The nearest whole number of stripes for indicating 78 r.p.m. is 77; for 45 r.p.m. it is 133, and for 33⅓ r.p.m. 180. The general formula is

$$\frac{\text{supply frequency} \times 120}{\text{turntable r.p.m.}}$$

As for the pick-up, the requirements are identical with those for "hi-fi"—level response over as wide a frequency band as possible, and absence of non-linearity. Most high-fidelity types are good in these respects, but give a small output. While it is easy to retain the accuracy of the record's frequency calibration, the amplitude calibration is affected by the characteristic of the pick-up and of any amplifier that may be necessary for increasing the output. Although the accuracy of measurements that are made with it is not thereby upset if properly carried out (see Sec. 11.6) a lot of time taken over adjustments can be saved if the output is substantially uniform over the whole band of frequencies. Whatever stylus is used it is most important that it be suited to the record in use, and is renewed before wear is appreciable.

The gramophone has its limitations, of course: the records wear out or get broken; it requires more manipulation than an all-electric source; it is unsuitable for long runs; the output is neither very large nor very uniform; and the range of frequencies is limited. Not only does the amplitude obtained even from a particular groove depend on the type and state of needle at the time, but so does the waveform. In fact, the only really reliable characteristic (given a known and constant speed of turntable rotation) is the fundamental frequency.

A magnetic tape recorder, though it would hardly be chosen solely as a cheap signal source, can sometimes be put to good use in the lab., for repeating signals as often as desired. Standard test tapes are available; see Sec. 14.38. Recorder characteristics at the top end of the a.f. band are generally not very good, and output is liable to fluctuate slightly; but continuity of recording is provided up to half an hour or more. And—although this has nothing to do with signal supply—a tape recorder is a useful unpaid assistant for taking down readings during experiments, especially those requiring concentration and speed.

4.15. OSCILLATORS: GENERAL REQUIREMENTS

All-electric signal sources consist of valve or transistor oscillators in more or less elaborated forms. In dealing with them it will be well to keep in mind other laboratory uses, of which there are chiefly two: as standards of frequency, and for measurements of resistance and reactance, etc. These are considered more fully in Chapters 6 and 9 respectively. In most applications it is necessary or desirable for the oscillator to have good waveform and constancy of frequency and amplitude. Unless properly designed it is likely to be more or less deficient in all of these respects, so the essential principles will now be reviewed.

In all of the various types of oscillator, considered individually in Secs. 4.18 to 4.23, the function of the valve or transistor is to produce the equivalent of negative resistance, with which to cancel the positive resistance of the oscillatory or tuning circuit. Usually this negative resistance is voltage controlled (point-contact transistors provide an exception) so is effectively in parallel with the tuning circuit, the positive resistance of which is the so-called dynamic resistance of that circuit. When two resistances are connected in parallel it is the *smaller* of them that dominates the partnership; so a high negative resistance is not so effective as a low one. Thus if a tuned circuit has a dynamic resistance of 50 kΩ, the negative resistance needed to cause oscillation would have to be 50 kΩ or *less*; if it were -60 kΩ the resultant would be $-60 \times 50/(50-60) = +300$ kΩ, its effect therefore being to increase the dynamic resistance and so reduce the damping, but not sufficiently to maintain continuous oscillation. The greater the negative resistance, the smaller its effect. This confusing inverted relationship will be avoided from now on by dealing in negative *conductance*. The more of it the valve provides, the greater the tendency to oscillation.

The key fact is that the amplitude of oscillation remains constant only when there is an exact balance between positive and negative conductance. If the combination is positive, oscillations die out; if it is negative, they grow. To make sure that they keep going it is necessary to provide an ample margin on the negative side; especially if the frequency is to be varied, because that varies the positive conductance. What happens, then, is that the amplitude of oscillation grows until the margin is reduced to zero by changes in characteristics, such as grid current that increases the positive conductance of a valve or a bottom bend that diminishes the negative conductance. If there is any change in the balance, due perhaps to tuning to a different frequency, or varying supply voltages or valve characteristics, the amplitude of oscillation has to change to restore it. In laboratory work, where the amplitude usually has to be known, or at least constant, this is inconvenient, necessitating monitoring and frequent adjustment.

Next, the non-linearity needed to remove the surplus negative conductance distorts the waveform. If in order to maintain oscillation over a large range of positive conductance, a lot of surplus has to be provided, there will be a lot of distortion. And as the amplitude varies, so will the distortion. This may not be very troublesome at radio frequencies, because it is easy to provide high-Q r.f. circuits to reduce the harmonics relative to the fundamental, and in any case their presence is often no disadvantage and may even be very useful. But in a.f. sources it is both more desirable and more difficult to achieve pure waveform.

Lastly, increasing amplitude and distortion are accompanied by increasing departure from the initial frequency of oscillation. The reasons for this are complicated, and there is a large literature on the subject; see, for example, *Theory and Design of Valve Oscillators*, by H. A. Thomas (Chapman & Hall). The standard of frequency stability expected nowadays is very high, so this effect, though small, is important.

The three main characteristics—frequency, waveform and amplitude —are thus closely bound up together, so in Sec. 4.24 special attention will be devoted to the means by which amplitude is limited.

Ref.: *Vacuum Tube Oscillators*, by W. A. Edson (Chapman & Hall, 1953).

4.16. REDUCTION OF HARMONICS BY *LC* CIRCUIT

Before going on to consider particular types of oscillator it may be as well to enlarge a little on one point in the last section—that tuned circuits greatly reduce distortion. How greatly is often not realized, although it follows from elementary theory of resonance. Suppose a

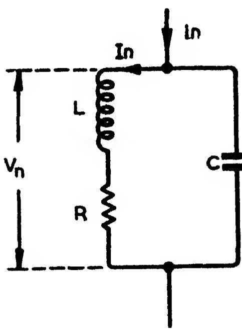


Fig. 4.16—Here i_n, I_n and V_n denote currents and voltage at the n*th harmonic of the frequency to which the circuit is tuned*

simple parallel tuned circuit (Fig. 4.16) is fed with currents of various frequencies. The ingoing current of the fundamental frequency (i.e., the one to which LC is tuned) is denoted by i_1; the second harmonic by i_2; and so on. The resulting currents flowing through L are called I_1, I_2, etc., and corresponding voltages across the tuned circuit are V_1, V_2, etc. Q is assumed to be at least 10, so $V_1 \simeq I_1\,\omega L$, etc. As is well known, $I_1 = Qi_1$, so the fundamental is magnified by a factor usually of the order of 100; but the harmonics are actually reduced, so relative to the magnified fundamental are very small indeed, as in Table 4.1, where i_1, i_2, etc., are all assumed to be equal.

The figures given in this table can be regarded as harmonic-reducing factors. For instance, if Q were 100, the fundamental voltage across the tuned circuit would be 150 times the second harmonic voltage, 267 times the third harmonic, etc. But this is on the basis of the harmonic currents fed in all being equal to the fundamental, which is an extreme of distortion that would not usually occur. If the maintaining current had, say, 30% second harmonic, the voltage across LC would have 30/150 = 0·2% second harmonic.

When the output of an oscillator is taken from across a high-Q tuned circuit, therefore, its waveform can never be very bad; but when taken

Table 4.1

Harmonic (n)	$\frac{I_1}{I_n}$	$\frac{V_1}{V_n}$
1 (fundamental)	1	1
2	$3Q$	$3Q/2$
3	$8Q$	$8Q/3$
4	$15Q$	$15Q/4$
5	$24Q$	$24Q/5$
6	$35Q$	$35Q/6$
n large	appr. n^2Q	appr. nQ

from elsewhere it is unlikely to be good, unless great care is taken to control the amplitude of oscillation.

If the tuned circuit is tapped, so that one branch has reactance of both kinds, the calculation is more complicated* and the reducing factor does not steadily increase with the order of the harmonic. In general, harmonics are reduced more if the tapping is on the capacitive side.

4.17. VALVES AND TRANSISTORS COMPARED

So far as oscillator design is concerned, the main differences are:

(1) Transistors have a very low input impedance.

(2) Manufacturers' tolerances on transistor characteristics are usually wider than on valve characteristics, although those are wider than on most other components.

(3) In general the gain (amplification) of a transistor begins to fall with increasing frequency at a much lower frequency—at a fraction of the "cut-off frequency", f_α. This fall is accompanied by an internal phase shift.

(4) There is an uncontrolled component of collector current which increases very steeply with temperature, so that germanium (and to a less extent silicon) transistors are limited to lower working temperatures.

(5) Transistors are more easily damaged by overloads.

(6) Transistor collector-current/base-current curves have a top bend as well as a bottom bend, in the usable current range.

* *Radio Frequency Measurements* (2nd ed.), by E. B. Moullin (Griffin), p. 111.

(7) The base needs the same polarity, with respect to the emitter, as the collector.

(8) A transistor is not a one-way device; the output circuit reacts on the input even at low frequencies.

For some of these reasons, laboratory oscillators (and other instruments) which are required to perform within close limits can more easily be designed around valves. For instance, to ensure oscillation under all conditions and with any transistor of the specified type, a larger margin of positive feedback must be provided, with the result that the waveform and the stability of amplitude and frequency are less certain. Alternatively, the circuit must be complicated by compensating devices.

On the other hand, transistors are much smaller, and so are their power requirements, and they need no warming-up period; so where size and portability and readiness are important their disadvantages may well be tolerated or overcome. In this, the amateur is better off than the manufacturer, who has to make all his instruments conform to specification using any transistor bearing the specified type number.

For most valve circuits there are transistor circuits which look the same except for base bias arrangements, but it should be noted that their operating conditions usually differ considerably.

4.18. FEEDBACK *LC* OSCILLATORS

Most oscillators consist of an amplifier with frequency-discriminating positive feedback. To maintain oscillation at a desired frequency the loss in the feedback circuit at that frequency must not be greater than the gain in the amplifier, and the total phase shift round the amplifier-and-feedback loop must be zero. In the usual type of single-valve or transistor amplifier (grid or base input; anode or collector output) there is a phase reversal, so the feedback circuit has to reverse it back again. Most commonly the frequency-discriminating or tuning device is an inductance-capacitance (*LC*) combination. The Hartley and Colpitts circuits are the two best-known varieties in this class, and they can be used as the basis of laboratory oscillators, especially for radio or high audio frequencies. *LC* oscillators with more than one stage in the amplifier are considered in Sec. 4.22.

To ensure frequency stability, the first necessity is to use stable coils and capacitors for the tuning circuit. The next is to reduce the influence of the maintaining valve on the frequency. Thirdly, things must be so arranged that the use made of the oscillator does not react in any way on the frequency or amplitude of oscillation.

The design of stable coils and capacitors is considered in Chapter 6.

To reduce the undesirable influence of the valve, it should be tapped across as little of the tuning circuit as possible. This tends to reduce the feedback below the point needed to maintain oscillation; so to counteract such a tendency it is desirable for the Q of the circuit and the amplification of the valve to be as high as possible. The coupling of the circuit to the valve should then be reduced to the minimum needed for oscillation. If this can be done in such a way as to make the coupling at higher (i.e., harmonic) frequencies less still, the maximum benefit will be derived. Fig. 4.17 shows an example of a frequency-stable type of oscillator embodying these ideas. It is sometimes called the Clapp oscillator, because it was described by J. K. Clapp (*Proc. I.R.E.*, March 1948); but it appears to have been devised earlier by G. G. Gouriet, though he wrote later (*W.E.*, April 1950). Although some of the original claims for this circuit were modified by discussion

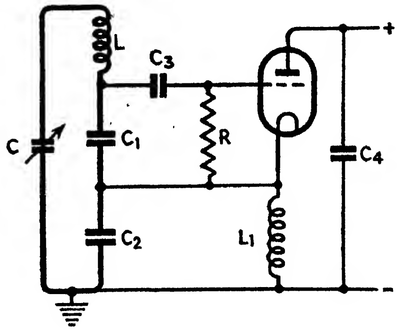


Fig. 4.17—Modification of the Colpitts oscillator circuit to give good frequency stability

(*W.E.*, May 1955, p. 141; Sept. 1955, p. 254–5; Jan. 1956, p. 53) in which emphasis was placed chiefly on the importance of high *Q*, in practice it does give very good stability (*W.W.*, Sept. 1957, p. 443).

The tuning circuit is drawn in heavy line, LC being the main frequency-determining components, and the capacitance of C_1 and C_2 is made as large as will offer only enough reactance to couple the valve and so have relatively little influence on frequency. R and C_3 are the usual grid leak and capacitor, L_1 is a r.f. choke to provide a conductive path to the cathode, and C_4 a by-pass to hold the anode at zero signal potential. The valve can of course be a tetrode or pentode.

A practical disadvantage is that unless C_1 and C_2 are varied along with C the system oscillates too fiercely at the maximum setting of C and not at all at the minimum, and the fact that C_1 and C_2 are of the order of 10 nF makes it more than usually inconvenient to vary them. So this circuit is chiefly of interest where the frequency is fixed, or variable only within narrow limits. More convenient for variable frequency is the corresponding Hartley circuit, in which the tapping down is on the inductor.

Because L and C in Fig. 4.17 are the main frequency-determining quantities, this circuit is sometimes considered to be series-tuned, but actually L and C cannot be in series resonance at the working frequency; they are together equivalent to an inductance whose value varies very rapidly with frequency, in parallel resonance with C_1 and C_2. In a crystal-controlled oscillator (Sec. 6.14) the crystal works in the same way, being equivalent to a very large fixed L in series with a very small fixed C, or alternatively a comparatively small variable inductance in tune with the parallel capacitance of the crystal and holder. Seen this way, it is no exception to the above rule that a high C/L ratio helps to stabilize frequency. This matter is discussed in "Series or Parallel?" and "L/C Ratio" (*W.W.*, Aug. and Sept. 1952).

Guiding principles in designing positive-feedback LC valve oscillators are to avoid grid current as far as possible, to employ valves of high resistance and high mutual conductance, to use oscillatory circuits of high Q and high C/L ratio, and minimum coupling between oscillatory circuit and valve.

The only necessary circuit change on substituting a transistor for the valve in Fig. 4.17 is to return R to the collector instead of the emitter. An oscillator of this kind was found by the author to have a temperature coefficient of frequency equal to $50/10^6$ per °C, compared with

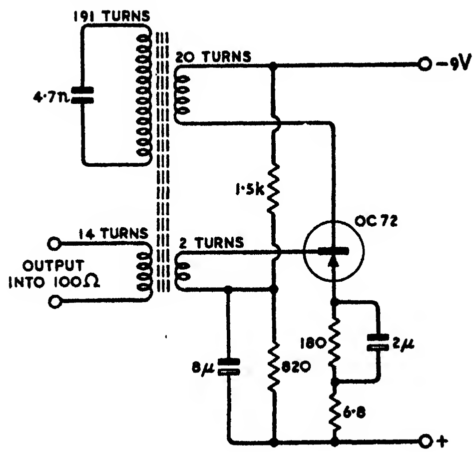


Fig. 4.18—Stable transistor oscillator circuit for 20 kc/s

$500/10^6$ using the same components in a conventional reaction-coil circuit (*W.W.*, Sept. 1957).

To ensure the desired emitter current in spite of change of temperature or transistor, the standard technique* is to insert a resistor in the emitter circuit (c.f. a cathode bias resistor) and—as this alone would bias the base with the wrong polarity—to apply in addition a suitable fixed bias by means of a low-resistance potential divider across the d.c. supply. Fig. 4.18 shows a practical 20 kc/s 5 mW oscillator, the design of which is discussed by J. F. Berry and L. E. Jansson (*M.T.C.*, Nov. 1958).

4.19. ELECTRON-COUPLED OSCILLATORS

The third condition for frequency stability was the isolation of the frequency-determining part of the circuit from load changes. One

* For a good detailed explanation see Chap. 4 of *Transistor Electronics*, by A. W. Lo and others (Prentice-Hall, Inc., 1955).

solution is to connect the oscillator to the output through a screened valve, which, as it need not necessarily amplify, is called a buffer stage. To avoid the need for the additional valve, various arrangements have been devised for one-way transfer of oscillation through the oscillator valve itself by electron coupling.

A feature of many of the circuits so named is that, as in Fig. 4.17, the cathode is not kept at constant potential but is tapped up the oscillatory circuit, though more usually on the coil than the capacitor side of it, as in Fig. 4.19. But this feature is no essential part of electron-

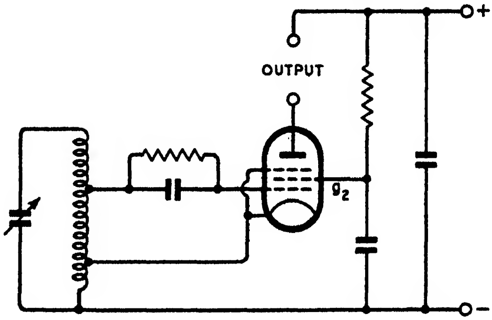


Fig. 4.19—Electron-coupled pentode oscillator circuit, in which g_2 is used as the oscillator anode, and the usual anode for output

coupled circuits in general. The original reason for it was to enable the screen grid (g_2) to be held at constant potential so as to exclude capacitive coupling between it and the anode. If a pentode is used, however, this method is not necessary, as the suppressor grid serves as a screen, and one is free to use conventional earthed-cathode circuits. It is interesting to note that in the original paper on electron coupling by J. B. Dow (*Proc. I.R.E.*, Dec. 1931), earthed-cathode circuits only are shown, even with tetrodes, whose capacitance coupling had to be " neutralized ". The merit of the tetrode oscillator, explained by Dow, is that by taking the g_2 feed voltage from a selected point on a potential divider across the anode supply it is possible to balance out the effect of varying supply voltage on frequency. This advantage is not obtainable with a pentode. Whether the cathode is earthed or not, there is a definite advantage in tapping it near the screen-grid end of the tuned circuit.*

The waveform of the signal in the anode circuit is usually far from pure, and either fundamental or harmonic frequencies can be selected by suitably tuning this circuit.

4.20. THE DYNATRON

A disadvantage of the class of oscillator circuit so far considered is that the maintaining amplifier has to be connected to LC at at least

* " Electron-coupled Oscillators ", *W.W.*, Dec. 1952.

three points. The dynatron is one of several maintaining devices that need be connected at only two points, normally the common terminals of L and C. Almost any screen-grid tetrode with the anode fed at a substantially lower voltage than g_2 will work as a dynatron. For detailed information on its applications and mode of operation, see *W.E.*, Oct. 1933.

The action of the dynatron depends on secondary emission from g_2. The curves in Fig. 4.20 show that when the anode voltage V_a is between the limits of about 10 to 90% of V_{g2} the slope of the anode current (I_a) curve, which represents the anode a.c. conductance, is downwards, which means that the conductance is negative, and any tuned circuit connected as in Fig. 4.21 will oscillate if its own conductance is less. As the curves show, the negative conductance can be varied by either V_{g2} or V_{g1}. Usually it is convenient to fix V_{g2} at the lowest that will give the required negative conductance and sufficient amplitude of oscillation, and to use V_{g1} for reducing the conductance until oscillation is only just maintained. Under these conditions oscillation sweeps

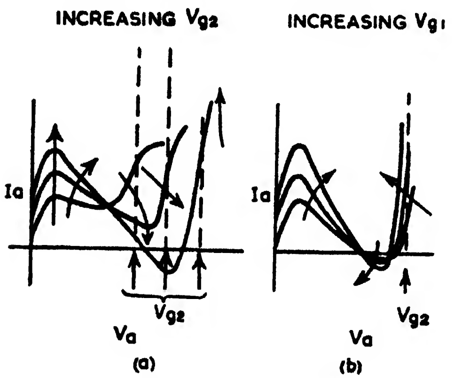


Fig 4.20 — Typical anode-current/anode-voltage curves of dynatron, showing the effects of varying g_2 voltage (a) and g_1 voltage (b). V_{g2} is marked along the anode voltage scale

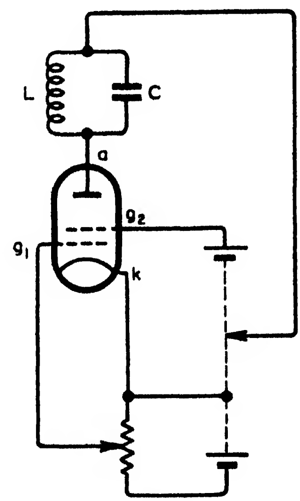


Fig. 4.21—Basic dynatron circuit

over the almost straight downward slope and the waveform is exceedingly pure. Suitable voltages are V_{g2} 100, V_a about 20 for small amplitudes and 50 for maximum, and V_{g1} variable 0 to −8. With zero bias the negative conductance of a good dynatron goes up to about 170 micromhos (resistance = −6 kΩ), capable of setting even heavily damped circuits into oscillation; but when run like this there is a risk of the dynatron properties deteriorating fairly quickly. Unlike other negative conductors, such as those described next, the dynatron's negative conductance is effective at z.f. An important practical consequence is that any potential divider used to tap off the anode voltage must be considerably lower in resistance than the negative resistance of the dynatron.

The advantages of the dynatron are its ability to set up oscillation in a simple two-terminal circuit, which may even be screened and inaccessible; the ease and precision of control; the straightness of its working characteristic; and the frequency-stability of its oscillations, provided that they do not sweep beyond this working slope. It lends itself particularly to automatic amplitude control (Sec. 4.25), because the control element g_1 forms no part of the oscillating circuit, and a very large control is exercised by a small change of voltage.

Its disadvantages are two: the secondary emission on which it depends is a fortuitous property, varying considerably among samples of the same type of valve; and the most suitable types are obsolete. But G. A. Hay has shown, in a valuable investigation (*W.W.*, Sept. 1944), that a number of types of pentode can be substituted, by the simple expedient of "commoning" g_2 and g_3. Modern pentodes, however, do not have the top-cap anode which was one of the advantages of the original dynatron for measuring purposes (Sec. 9.32).

The same investigator has shown (*W.E.*, Nov. 1946) how the dynatron can be used for measurements up to 50 Mc/s or higher. It appears that the inaccuracies and irregular results experienced with the dynatron at very high frequencies were not due, as had been supposed, to transit-time effects in the valve, but to resonances in the external connections. By keeping the r.f. leads as short as possible, and using chokes to exclude r.f. currents from all leads in which they have no business, measurements can be made accurately well into the v.h.f. band. Above 50 Mc/s, internal resonances are liable to give trouble in valves of normal construction.

For frequencies far above the capabilities of the dynatron, the recent tunnel diodes are likely to prove useful two-terminal negative resistors.

4.21. THE TRANSITRON

Another device capable of producing a negative conductance between a pair of terminals is the transitron, first described by E. W. Herold (*Proc. I.R.E.*, Oct. 1935) but so called by Brunetti, whose paper on it (*Proc. I.R.E.*, Dec. 1937) is worth studying also for its explanation of "average negative resistance". LC is connected in the g_2 circuit (Fig. 4.22) and the changes of voltage across it have to be passed on to g_3 in order to generate the negative resistance or conductance. The difference in feed voltage between g_2 and g_3 must be maintained, usually by a blocking capacitor C_1, g_3 being tied to the required voltage—preferably slightly negative—through R. These appendages make the circuit less ideally simple than the dynatron, and increase the admittance across LC—a disadvantage when measuring high dynamic resistances (Sec. 9.33). For most other purposes, however, this need not be troublesome, especially if C_1 is a physically small ceramic capacitor connected close up to the electrodes, and R is of the order of 1 MΩ.

As with the dynatron, g_1 is a convenient throttle for adjusting the value of negative conductance. V_a and V_{g2} giving best negative conductance characteristics depend on the type of valve, but for long life should not be needlessly high. Articles on the practical use of the transitron as an oscillator include those by A. G. Chambers (*W.W.*, March and April 1943) and F. P. Williams (*W.W.*, Aug. 1944). In the latter it is shown that for satisfactory operation g_3 should be provided with sufficient bias to prevent it from being swung positive by g_2, and the recommended control system for the EF50 valve is repeated at Fig. 4.23.

A. G. Bogle has shown (*E. & R.E.*, May 1957) that the negative resistance provided by a transitron is (like a dynatron) inversely

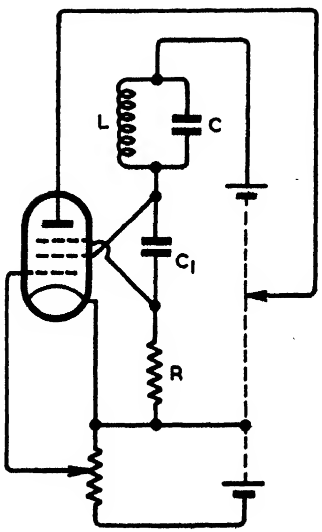


Fig. 4.22—Basic transitron circuit

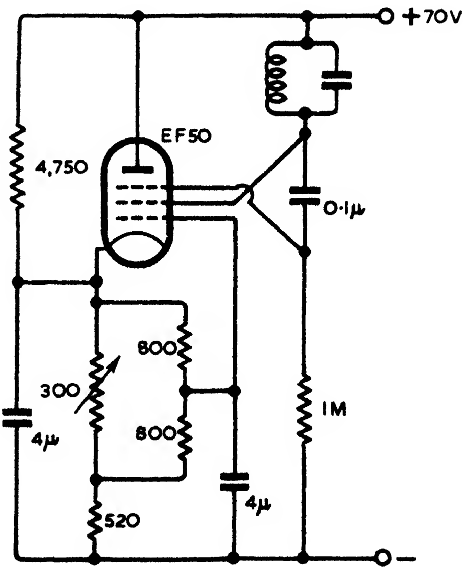


Fig. 4.23—Practical transitron circuit

proportional to its cathode current, within about 10%, and that this property can be utilized for measuring positive r.f. resistance.

4.22. AMPLIFIER TWO-TERMINAL *LC* OSCILLATORS

The need for more than two connections to the tuned circuit in the feedback oscillators referred to in Sec. 4.18 is to obtain the phase reversal to cancel the reversal in the single-valve common-cathode amplifier. If two valves are used this is not necessary, and there are other advantages, such as greater flexibility in design and a larger obtainable negative conductance. Incidentally, the cost of a twin triode is about the same as that of a single pentode.

One variety, the basic circuit of which is shown in Fig. 4.24, has been described by F. Butler (*W.E.*, Nov. 1944). It can be regarded as developing a negative conductance between the LC terminals. V_1 is a common-grid amplifier, a type which has a low input resistance ($\frac{r_a + R_L}{\mu + 1}$, where R_L is the anode load resistance) so has to be driven by the cathode follower V_2. In a practical version of the circuit given by Butler, R_2 is 10 kΩ, but as this is too much for bias purposes the grids of both valves are tied through " leaks " to a tapping 500 Ω from the cathodes.

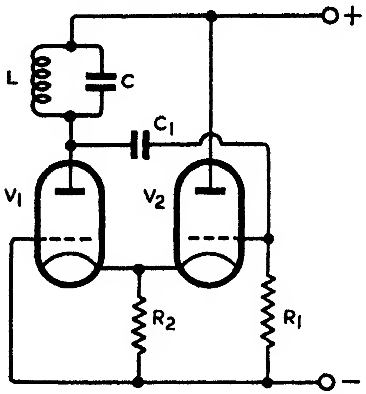


Fig. 4.24—Outline circuit of cathode-coupled oscillator

In a modification of this cathode-coupled oscillator that has received a good deal of attention in America, R_1 and LC are interchanged. Its remarkable adaptability to a variety of laboratory uses, with great simplicity, has been brought out by K. A. Pullen, Junr. (*Proc. I.R.E.* June 1946).*

The Butler version is preferable if the oscillator is required as an alternative to the dynatron or transitron in r.f. resistance measurements, for the admittance in parallel with LC can be kept quite low—especially if pentodes are used, with g_2 tied to cathode by capacitor. On the other hand, as a variable-frequency signal source the alternative version puts LC where it is more conveniently earthed, and the fact that R_1 is necessarily of the kΩ rather than the MΩ order helps to stabilize the amplitude of oscillation. If the output is taken from across R_2 the reaction on frequency is comparatively small, but it should be noted that unless the amplitude of oscillation is restricted the waveform

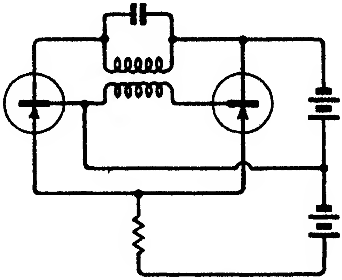
* Other helpful references are:
" Two-terminal Oscillator ", by M. G. Crosby. *Electronics*, May 1946.
" Cathode-coupled Negative Resistance Circuit ", by P. G. Sulzer. *Proc. I.R.E.*, Aug. 1948.
" Frequency and Amplitude Stability of the Cathode-coupled Oscillator ", by P. G. Sulzer. *Proc. I.R.E.*, May 1950.

across R_2 may be very distorted (Sec. 4.19). The greatest negative conductance is obtained if R_2, with usual valves, is of the order of 1 kΩ.

In yet another version of this circuit, a calculable negative resistance is developed in the common cathode lead (H. J. Reich, *Proc. I.R.E.*, Feb. 1955).

The performance of what J. R. Tillman classifies as a precision negative resistance has been detailed by him (*W.E.*, Dec. 1947). It

Fig. 4.25 — Emitter-coupled transistor oscillator needing no tappings on tuned circuit



consists of a high-gain amplifier, incorporating negative feedback, with sufficient positive feedback via the *LC* circuit to keep it in oscillation. High stability of frequency is obtained, and amplitude is unusually constant. Its superiority to the dynatron and transitron as regards its negative conductance being largely independent of changes of supply voltages and valves is shown by the same author (*W.E.*, Jan. 1945). (See also " The Principles and Design of Valve Oscillators ", by A. C. Lynch and J. R. Tillman. *E.E.*, Feb. and March 1945).

Pure waveform ($<0{\cdot}1\%$ harmonic distortion) and readily-calculable performance, largely independent of variations in transistor parameters, are claimed by P. J. Baxandall (*Proc. I.E.E.*, Vol. 106, Part B, Supp. 16, May 1959) for a transistor oscillator, Fig. 4.25, which is essentially the same as Fig. 4.24 but with inductive instead of direct coupling. Other two-valve transistor oscillators in the same paper are capable of efficiencies up to 85%.

4.23. *RC* OSCILLATORS

Many of the oscillators mentioned so far can be used satisfactorily at frequencies up to or including v.h.f., if the appropriate *L* and *C* are connected to the negative resistance terminals. In another class of oscillator circuit the frequency-determining elements in the feedback path are resistance and capacitance (*RC*). Although such oscillators can be made to work at r.f., their advantages increase toward the low-frequency end, especially below 100 c/s, because at such frequencies air-cored coils have excessive bulk and resistance, and iron-cored coils tend to cause poor waveform and inconstant frequency.

As with *LC* oscillators, there are single-valve and multi-valve varieties, and in the single-valve variety the *RC* circuit must shift the phase 180° without losing more voltage than the valve gains. The greatest phase shift given by one capacitor and one resistor is less than 90°, so at least three stages are needed, as shown in outline in Fig. 4.26.

The frequency at which the phase shift is 180°, and therefore (neglecting the effect of valve and load resistances) the frequency of oscillation is

$$f_0 = \frac{1}{2\pi CR\sqrt{6}}$$

and the voltage amplification needed to make good the loss in the

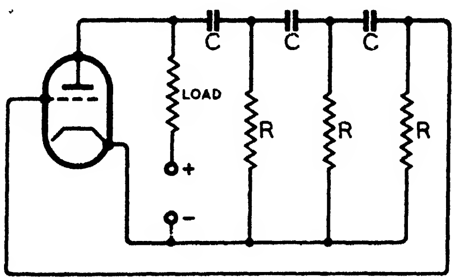


Fig. 4.26—Outline circuit of a simple form of phase-shift oscillator. The frequency of oscillation depends on the values of R and C

three-stage *CR* network is 29, or just over 29 dB. With a four-stage network it is 25·3 dB.

R and C can be interchanged, in which case

$$f_0 = \frac{\sqrt{6}}{2\pi RC}$$

i.e., 6 times as great. This arrangement is therefore preferred for high-frequency oscillators, as explained by G. W. Holbrook (*E.E.* Dec. 1953). The limiting frequency is about 5 Mc/s, but in practice can hardly reach that. Formulae for the somewhat lower values of f_0 when valve and load resistances are taken into account are given by Holbrook, and also by W. C. Vaughan in a very comprehensive treatment of the single-valve phase-shift oscillator (*W.E.*, Dec. 1949).

It is worth noting that as the *CR* type acts as a high-pass filter the least distorted waveform is obtained at the input (anode) end, whereas the opposite applies to the *RC* type.

Since the frequency-determining network lacks the large flywheel effect of a high-*Q* *LC* circuit,* over-oscillation results in much greater deterioration of waveform than in a comparable *LC* oscillator; but

* "The Equivalent *Q* of *RC* Networks", by D. A. H. Brown. *E.E.*, July and Sept. 1953.

provided the gain of the valve is adjusted so as barely to offset the loss through the network a very good sine waveform can be obtained. An example with <0·02% harmonic distortion is included in Sec. 4.28. To ensure this, automatic control (Sec. 4.25) is practically indispensable.

In an alternative type of phase-shift oscillator (W. Fraser, *E.E.*, May 1956) two 90° shifters of the Fig. 4.27 type are sandwiched between three valves. Frequency is varied over a 10 : 1 range by R and in five decade steps by C, giving the wide range of 1 c/s to 100 kc/s, which is especially useful for testing a.f. amplifiers having negative feedback.

If two valves are used, the required phase shift in the feedback path is zero. The usual device for obtaining this at a single frequency is a potential divider, one arm made up of C and R in series and the other of C and R in parallel, as in Fig. 4.28, which is the *RC* counterpart of the cathode-coupled *LC* oscillator (Fig. 4.22). The only frequency at which the output of this potential divider is in phase with the input (and, incidentally, is at its maximum) is

$$f_0 = \frac{1}{2\pi\sqrt{(R_1R_2C_1C_2)}}$$

in which R_1 includes the valve and load resistances in parallel with one another; and the required voltage gain is $1 + R_1/R_2 + C_2/C_1$. When,

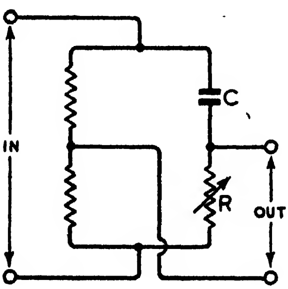


Fig. 4.27 — Variable-frequency 90° phase shift circuit, two of which are used in a multi-valve oscillator

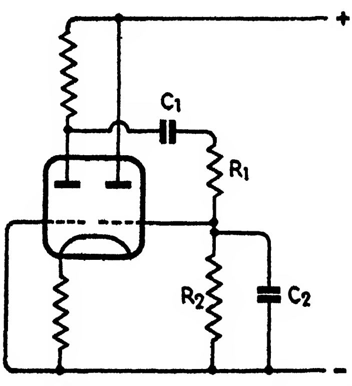


Fig. 4.28—A particularly simple type of RC oscillator

as is convenient for ganged frequency control, $R_1 = R_2$ ($=R$) and $C_1 = C_2$ ($=C$), this frequency is inversely proportional to R and C, and the output voltage is one third of the input (attenuation = 9½ dB). If therefore the gain of the amplifier is 9½ dB and its phase shift zero, oscillation will be maintained at frequency f_0.

Any phase shift in the amplifier necessitates a corresponding shift in the *RC* network, and consequently a departure from the frequency given in the equation. To minimize amplifier phase shift and so stabilize the frequency, negative feedback is usually introduced. A commonly-used method is to feed the input in opposite polarity with a proportion of the output, tapped off by a potential divider (R_3R_4 in Fig. 4.29).

Another way of looking at the circuit is to consider it as a bridge (actually known as the Wien bridge, Sec. 7.19) which would be balanced if $R_3 = 2R_4$, because both of the " detector " points would be at the

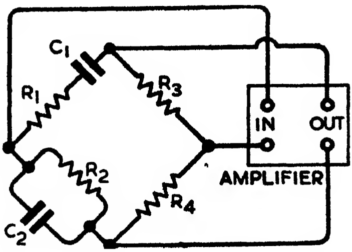


Fig. 4.29—Functional diagram of bridge-controlled RC oscillator

same potential. In other words, the attenuation of the network would be infinitely large. By lowering the tapping on R_3R_4 the attenuation is reduced, until the loop gain of the amplifier is sufficient to cause oscillation.

If the gain of the amplifier itself is made very much larger than 9½ dB, the necessary shift in the R_3R_4 tapping is small, so that the $R_3 : R_4$ ratio becomes extremely effective as an oscillation control. By using a lamp for R_4, or, better still, a thermistor for R_3, oscillation is automatically controlled. A very satisfactory audio oscillator can be designed on these lines, and practical examples are described in Sec. 4.28. For more about their theory and design see D. E. D. Hickman (*W.W.*, Dec. 1959).

Although the usual method of frequency control is to vary R (or C) continuously over a 10 : 1 range and C (or R) in decade steps, there are obvious advantages in obtaining the full range in one sweep of the control. One method of obtaining a 1,000 : 1 frequency range, by means of a number of components permanently connected instead of switched, was described briefly (with reference to original paper) in *E. & R.E.*, July 1957, p. 272. The same principle is carried further, but more simply, using a single resistor with distributed capacitance, by C. G. Mayo and J. W. Head (*E. & R.E.*, Nov. 1958). And 20 c/s to 3 Mc/s in one sweep is claimed by F. S. Anderson (*Proc. I.R.E.*, Aug. 1951), with a four-valve circuit using a ganged pair of linear potentiometers.

For fixed-frequency oscillators there is no objection to making R_1 and R_2, and C_1 and C_2, unequal; in fact in an article " Optimum Conditions for RC Oscillators " (*Electronics*, Feb. 1948), H. A. Whale has shown that the greater C_2/C_1 and R_1/R_2 the better the frequency stability.

The design of *RC* oscillators using transistors is influenced by their internal phase shifts (which upset calculated frequencies and limit the maximum more severely than with valves) and especially their low input impedance, whereby the *RC* network should be considered as an attenuator of current rather than voltage. The phase shifters described for valves should therefore be turned back to front, and in general their impedances should be lower. Design is discussed by D. E. Hooper and A. E. Jackets (*E.E.*, Aug. 1956), and by M. K. Achuthan (*E. & R.E.*, Aug. 1957).

4.24. AMPLITUDE STABILIZATION

As explained in Sec. 4.15, the key to desirable oscillator characteristics is control of amplitude. A margin of negative conductance is needed to make sure of oscillation, and the greater the possible variations in tuning circuit, supply voltages and load, the greater the margin

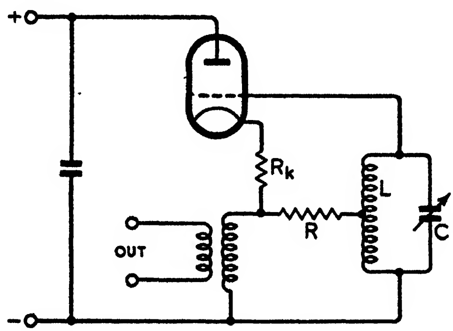


Fig. 4.30—Stable oscillator circuit, in which amplitude is controlled by R

that must be allowed, and the greater the variations in amplitude, waveform and frequency are likely to be, unless amplitude is controlled.

One method is to adjust the feedback very carefully by hand. But the setting is generally far from stable, and almost invariably the control must be reset whenever the frequency is altered. If a definite amplitude is required it is practically essential to use an indicator to show when it is reached, and in taking many readings these adjustments waste time.

The working point on the valve or transistor characteristics should be chosen so that the margin is steadily reduced as the amplitude increases; if the reverse takes place there is a sudden jump to a greater amplitude. And the method of feedback control should preferably be one that does not cause acute non-linearity, or changes in reactance or phase angle that would affect the frequency.

One good method is by the resistor R in Fig. 4.30. It is discussed by T. Roddam (*W.W.*, Feb. 1954), who claims that this oscillator circuit is simple, stable, and pre-eminently " designable ". The tuning inductor L is centre-tapped, and separated from the maintaining valve by *R* of the order of 15 kΩ. R_k is to provide Class A bias if the output transformer primary (designed to divert not more than half the oscillatory power from LC) has insufficient resistance.

A transistor circuit of somewhat similar merits is mentioned in Sec. 4.22.

The use of negative feedback in Tillman's oscillator (mentioned in Sec. 4.22) ensures that up to a certain amplitude the system is closely linear and beyond that a sharp limiting action sets in, which keeps the amplitude reasonably constant. While this distorts the waveform in the amplifier, perhaps very severely, it must be remembered that the influence of this on the waveform in the high-*Q* tuned circuit itself is reduced by a factor of hundreds (Sec. 4.16).

A pre-set feedback control claimed to maintain constant amplitude and good waveform even in a *RC* oscillator is described by E. J. B. Willey (*W.W.*, June 1947).

4.25. AUTOMATIC AMPLITUDE CONTROL

To avoid the uncertainty of manual control, several automatic systems have been devised. In a sense every oscillator has automatic control, by the mechanism described in Sec. 4.15, but the term is used for those systems that aim to restrict oscillation to a practically constant amplitude within the most linear range of valve characteristics, notwithstanding large variations in conditions.

One system is an adaptation of a.g.c: the oscillatory output is rectified to provide a bias voltage which is used to control oscillation. The grid capacitor and leak in circuits of the type shown in Fig. 4.17 act in this way by biasing back the grid when the amplitude exceeds the initial bias. The technique of amplified a.g.c. can be applied to obtain a firmer control (see L. B. Arguimbau in *Proc. I.R.E.*, Jan. 1933).

In one commercially produced unit the rectified peak voltage from the oscillator is used to control the voltage of the power supply by the usual technique (Sec. 4.5) of comparison with a voltage reference (W. F. Byers, *G.R.E.*, April 1955).

In some types of oscillator these methods are very useful, but the control circuits tend to become rather elaborate, and there is difficulty in choosing filter time-constants that give effective smoothing of the control voltage over the working range of frequencies without " hunting "—an effect caused by the amplitude of oscillation alternately growing and then being choked back by the control.

Methods generally favoured employ resistors whose resistance varies with temperature and hence with the current passing through them. Ordinary metal-filament lamps behave in this way, and they have been used for controlling *RC* oscillators; e.g., R_4 in Fig. 4.29. Thermistors—resistors made of certain materials whose resistance falls very steeply with rising temperature (and therefore current)—can be obtained in a wide range of values suitable for almost any circuit; the action is

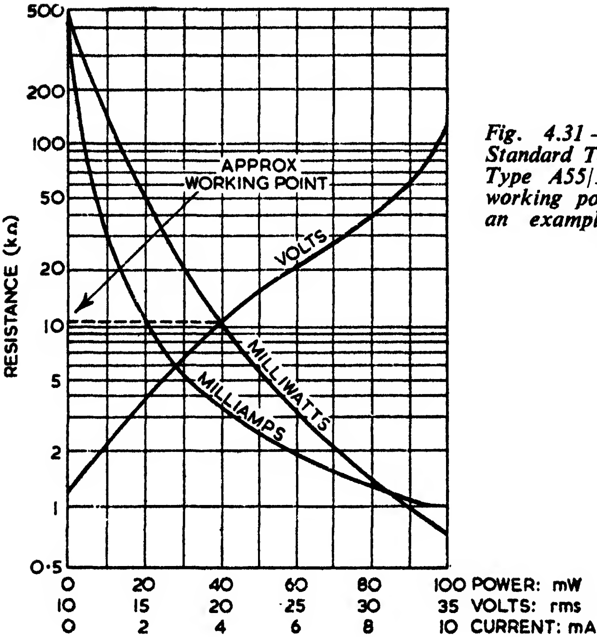


Fig. 4.31 — Characteristics of Standard Telephones thermistor, Type A55/13/100. A suitable working point is marked. For an example of its use, see Fig. 4.34

the opposite of that of lamps and is suitable for R_3 in Fig. 4.29. A practical example is given in Sec. 4.28. The properties and applications of thermistors are reviewed by K. R. Patrick (*E. & R.E.*, July 1958). Being directly sensitive to temperature, as well as indirectly to current, they can also be used to stabilize transistor oscillators against temperature effects (H. D. Polishuk, *E. & R.E.*, Oct. 1958).

Fig. 4.31 shows the characteristics of one type of thermistor. The general method of design is to choose a working point that gives a resistance suitable for the circuit and calls for an appreciable fraction of the maximum power. If one works close to the initial resistance the ambient temperature is likely to have too much effect on the amplitude of oscillation. A suitable working point is marked in Fig. 4.31, just over 10 kΩ, corresponding to a dissipation of 40 mW (20 V, 2 mA).

In a bridge or other system for magnifying the rate of control, such as that in Fig. 4.29, a thermistor can be made to hold the amplitude within very close limits, and does not distort the waveform. In spite of their effectiveness of control, thermistors are not prone to hunting, they absorb much less power than lamps, and are convenient and adaptable oscillation governors.

The various types of a.a.c. are compared by P. R. Aigrain and E. M. Williams in "Theory of Amplitude-stabilized Oscillators" (*Proc. I.R.E.*, Jan. 1948). Arguimbau's original paper on the rectified-feedback type has already been mentioned. The original paper on the bridge-stabilized type, by L. A. Meacham, is in *Proc. I.R.E.*, Oct. 1938 ; see also " Variable-frequency Bridge-type Frequency-stabilized Oscillators " by W. G. Shepherd and R. O. Wire (*Proc. I.R.E.*, June 1943), and " Thermistor-regulated Low-frequency Oscillator " by L. Fleming (*Electronics*, Oct. 1946). The design of a lamp-in-bridge controlled 20-kc/s oscillator of very high frequency stability is detailed by T. Roddam (*W.W.*, Aug. 1948).

4.26. THE BEAT-FREQUENCY SOURCE

For some purposes it is helpful to be able to cover a wider range of frequency with one sweep of the control than is possible with any of the forĕgoing types. The variable capacitor of an *LC* oscillator seldom gives a frequency range of more than about 3 : 1. In the *RC* oscillator, f_0 is proportional to $1/C$ (or $1/R$) rather than $1/\sqrt{C}$, so a

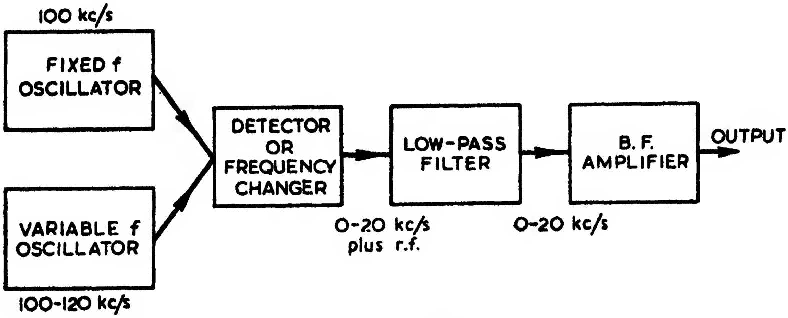


Fig. 4.32—Block diagram of beat-frequency a.f. source, with typical frequency values

range of 10 : 1 is easily attainable. Artifices for extending this to as much as 1,000 : 1 have been outpaced by the demand for 10,000 : 1 even for a.f. work (Sec. 4.27); and the modulation frequencies in television are much wider still—from zero up to several megacycles per second.

For very wide frequency range the usual type is the beat-frequency oscillator. It was probably suggested by the howl obtained with the old-time oscillating receiver when it beat with a received oscillation. The whole audible scale, and more, results from a very short easy movement of the tuning capacitor. In this the b.f. oscillator has a powerful attraction, so powerful that it may sometimes obscure the difficulties involved in cultivating the original howling receiver into a laboratory instrument of precision. So it may be as well to indicate some of the requirements, if only to deter thrifty readers from too light-heartedly setting out to make one. The essential stages are shown in Fig. 4.32, with typical frequencies for an a.f. source given as an example.

The first thing one notices about its prototype—the oscillating receiver—is that the very low frequencies are unobtainable because of the tendency for two oscillators to " pull in " when they are very close in frequency. To avoid this the oscillators must be very completely screened from one another. Yet their outputs have to be combined. The same problem is familiar in superhet receivers, where the two frequencies are separated by at least 450,000 c/s; but in a b.f. oscillator they must approach to 5 c/s, or preferably less, before pulling in. This does not necessarily mean that such a low frequency is wanted for testing; but if the two oscillators pull in at 5 c/s that fact implies poor waveform even at a considerably higher frequency.* Although some degree of success has been achieved in a simple form of instrument by combining the direct outputs of the oscillators in a frequency-changer valve, the standard method is to interpose buffer stages to prevent either oscillator from reacting on the other.

Waveform is particularly important if the source is intended for investigating amplifiers for small percentages of distortion, because it is obviously no good trying to do so if the test source itself is impure. Even for taking frequency-response curves a pure source is necessary. If the falling-off in response of some apparatus at, say, 50 c/s is being measured, and the nominal 50-cycle output of the signal put into it includes harmonics, to which the apparatus may be much more responsive, the result does not indicate the true 50-cycle response.

Apart from the distortion at low frequencies due to the tendency to pull in, there is distortion due to the detector, and to the beat-frequency amplifier that follows.

At least three systems of extracting the beat frequency have been adopted. In the first, oscillations of equal amplitude are applied to a square-law detector. If it is not exactly square-law, it will generate harmonics of the beat frequency. The amplitude of the beat is proportional to the product of the two ingoing amplitudes. In the second and commonest system, one oscillation is made strong enough to sweep up to the most linear range of a detector; the other oscillation is relatively small, say one-tenth, to keep within this linear range. The third, due to C. G. Mayo ("Beat-frequency Tone Source", *W.E.*, June 1952) is a modification of this, in which the larger oscillation is given a square waveform before combining with the smaller, a sine wave. For similar results, less input is required.

In both these linear systems the amplitude of the beat is determined by the smaller and not at all by the larger of the component oscillations; so in order to maintain a uniform output over the whole scale the smaller oscillation is made the fixed frequency, and the larger the variable. It is allowable, however, to vary the weaker oscillation over a narrow range of frequency, because the change in amplitude over such a range is generally negligible. The usefulness of doing so is that it permits a fine adjustment of frequency with a calibration that holds good at any setting of the main frequency dial. The main tuning capacitor can be shaped to give any desired scale shape (within reason). The C.C.I.R. international standard is logarithmic between 100 and 10,000 c/s and linear outside these limits.

* Actually 10% second-harmonic distortion at 50 c/s, 25% at 20 c/s, and so on. See "Distortion in Beat-frequency Sources", by C. G. Mayo. *W.E.*, April 1952.

If *both* oscillations include harmonics, harmonics of the beat frequency are produced by the detector; so a filter is usually included to make sure that at least one (the fixed, more easily) is of very pure waveform. Another reason for excluding harmonics in the component oscillations is that they are liable to cause spurious whistles at the top end of the frequency scale.

Distortion in the amplifier is minimized by careful design generally, by running the valves well below their maximum rated output, and by the use of negative feedback. To minimize distortion in the output stage it is generally necessary to ensure that the load impedance is reasonably close to the optimum for the stage. If the oscillator is used to feed a variety of loads, a well-designed multi-ratio transformer is almost essential. If the oscillator is not required to feed loads of less than about 100 Ω, it is possible to dispense with the transformer and reduce risk of distortion by adopting a cathode-follower output.

All trace of the component oscillations must be absent from the output of a b.f. source, or very misleading results might occur. A filter following the detector is necessary, and it must be well designed if it is not to start cutting the highest beat frequencies. The difficulty is that if, in order to simplify this problem, the component frequencies are made high, the pull-in and frequency-stability problems are increased. Assuming 100 kc/s for the fixed frequency, it is obvious that in order to maintain a certain constancy of the output frequency at, say, 100 c/s, the component frequencies must be 1,000 times as stable, unless, of course, they both shift equally in the same direction.

This problem is tackled by using very stable materials for the oscillator components; designing coils and capacitors so as to be unaffected by temperature; placing them well away from the heating components such as valves, or in heat-insulated compartments; arranging them symmetrically, to ensure equal changes in both oscillators; and making careful choice of oscillator circuit, as already discussed. It really is very difficult to design and make a b.f. source which will give an output that is constant in frequency within, say, 1 c/s from the time of switching on. The worst drift can be avoided by allowing a quarter of an hour or so for warming up, but often this is inconvenient.

Another necessity, if the instrument is mains-driven, is a very high standard of supply-current smoothing and absence of stray pick-up from transformers, etc. Quite a small hum content—say 1 % or even less—can be extremely annoying, not to say misleading, when testing an amplifier for non-linearity distortion.

Altogether then, a *really* good b.f. source cannot be bought cheaply or designed easily.

4.27. COMMERCIAL A.F. AND V.F. SOURCES

The Sullivan-Ryall b.f. source deserves special mention because of the amazing standard achieved—pull-in frequency below 0·1 c/s; stability from switching on, ±1 c/s per day, including temperature changes; harmonic content, less than 0·1% at 1 W output; maximum output, 15 W; amplitude constant to ±0·1 dB over whole scale and with ±5% supply voltage fluctuations; frequency accuracy, at least 0·2% ±0·5 c/s over the whole scale, 0 to 16 kc/s.

For most purposes, however, the *RC* oscillator has largely ousted the b.f. type, at least for audio frequencies. Although the formerly usual 20 c/s to 20 kc/s frequency coverage in three decade ranges does include all that are usually reckoned audible, most a.f. sources now go up to at least 100 kc/s and as far below 20 c/s as the makers consider practicable. This is necessary because of the tendency for the response of negative feedback amplifiers to peak outside the a.f. band. The Marconi TF 1370 covers the exceptionally wide range of 10 c/s to 10 Mc/s (sine waves) and up to 100 kc/s square waves

In some instruments, such as the Advance types H and J (10 c/s to 50 kc/s) and the Furzehill G.432 (25 c/s to 250 kc/s) the Wien bridge is continuously tuned by a gang capacitor and the ranges are obtained by switching resistors. This has the advantages that control of frequency is very smooth and the impedance of the bridge is constant over any one range, but of course it changes from range to range. The unavoidably low maximum capacitance leads to very high impedance on the lowest-frequency range—of the order of 10 MΩ—and the capacitor and other parts must be well screened to exclude hum. This is avoided by resistance tuning and switched capacitors, but the frequency then changes in very small steps so that it may not always be possible to set the control precisely to a given frequency. An advantage is that control is spread over about 300° instead of being confined to 180° as with variable capacitors.

In the Muirhead D-890 decade oscillator, an account of which appeared in *Muirhead Technique*, July 1959, the frequency is controlled by four 11-way switches, in steps of 1 c/s from 1 to 11,110 c/s, and there is a ×10 frequency multiplier.

The Wayne-Kerr S.121 oscillator combines both methods; the major frequency intervals—ten per decade—are switched, the numerical values being displayed in windows at each end of an interpolating scale, operated by a fine frequency control.

Still another frequency-control device, in the Furzehill G.432, enables that *RC* type to share an advantage usually possessed only by the b.f.o.—an incremental frequency scale, valid at any setting on the main scale. It is due to an additional phase-shift device as in Fig. 4.33. The conventional frequency control would be by C_4 and C_5, ganged. $C_1C_2C_3R_1$ is the additional phase shifter, controlled by C_3. To keep its scale correct at all settings of the main control, the latter adjusts C_1 also.

The Dawe Type 443 is an example of 20 c/s to 20 kc/s being covered in one sweep, by a Wien bridge modified by using *RC* networks in place of the usual resistors.

Type 421 by the same maker covers the same frequency band with the usual three switched ranges, but using transistors instead of valves.

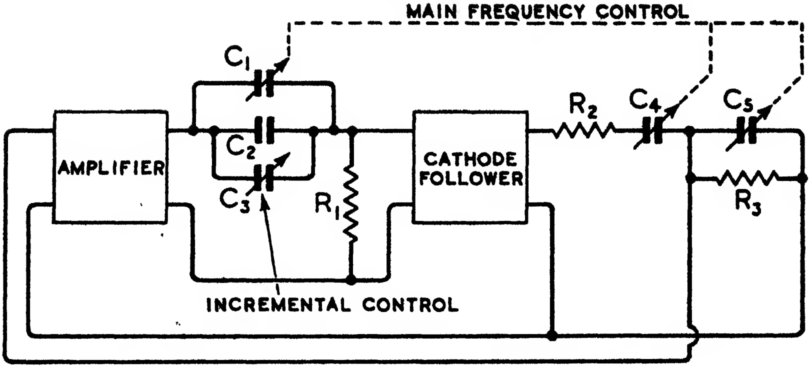


Fig. 4.33—Device for providing an incremental frequency scale in an RC type of oscillator—the Furzehill G.432

The low-impedance difficulty is overcome by feeding into push-pull emitter-follower stages, and amplitude stabilization is by a very low-power thermistor. The Levell Type TG150 is a battery-driven transistor oscillator covering 1·5 c/s to 150 kc/s in five ranges. Distortion less than 0·1 % is claimed.

For v.f. testing, Marconi Instruments TF.855A/1 b.f. oscillator covers 25 c/s to 12 Mc/s in three stages, and also provides 50 c/s to 150 kc/s square waves. The Wayne-Kerr Type 022B covers 10 kc/s to 10 Mc/s with sine waves having less than 1 % harmonic distortion. See also another kind of v.f. generator, the Pye PTC 1201/3, in Sec. 4.38.

To avoid transformer difficulties, the output stage of most of these oscillators is by cathode follower, but there is a tendency to use " single-ended push-pull " stages.

Low frequency (including very low frequency, not included here) sine-wave generators are reviewed in *E. & R.E.*, Dec. 1958.

4.28. CONSTRUCTIONAL A.F. SOURCES

A number of designs, suitable for the experimenter to make up himself, have been published. One by the author, although dated 1949, is still fairly typical except for its 20 c/s to 20 kc/s frequency range—and that can easily be extended. It continues to give very satisfactory results.

Fig. 4.34 is the circuit diagram. Frequency is controlled by a 80 + 80 kΩ ganged rheostat and pairs of switched capacitors. A close approximation to a logarithmic frequency scale is obtained by using graded rheostat elements known as semi-log or inverse semi-log, according to whether the control is of the rotating-dial fixed-pointer type or vice versa. By picking the capacitors carefully and building them up to exact decimal multiples of one another, one scale can be made to serve all ranges.

The first two valves form a two-stage amplifier, with feedback via a Wien bridge, and the third is a cathode-follower output stage. The

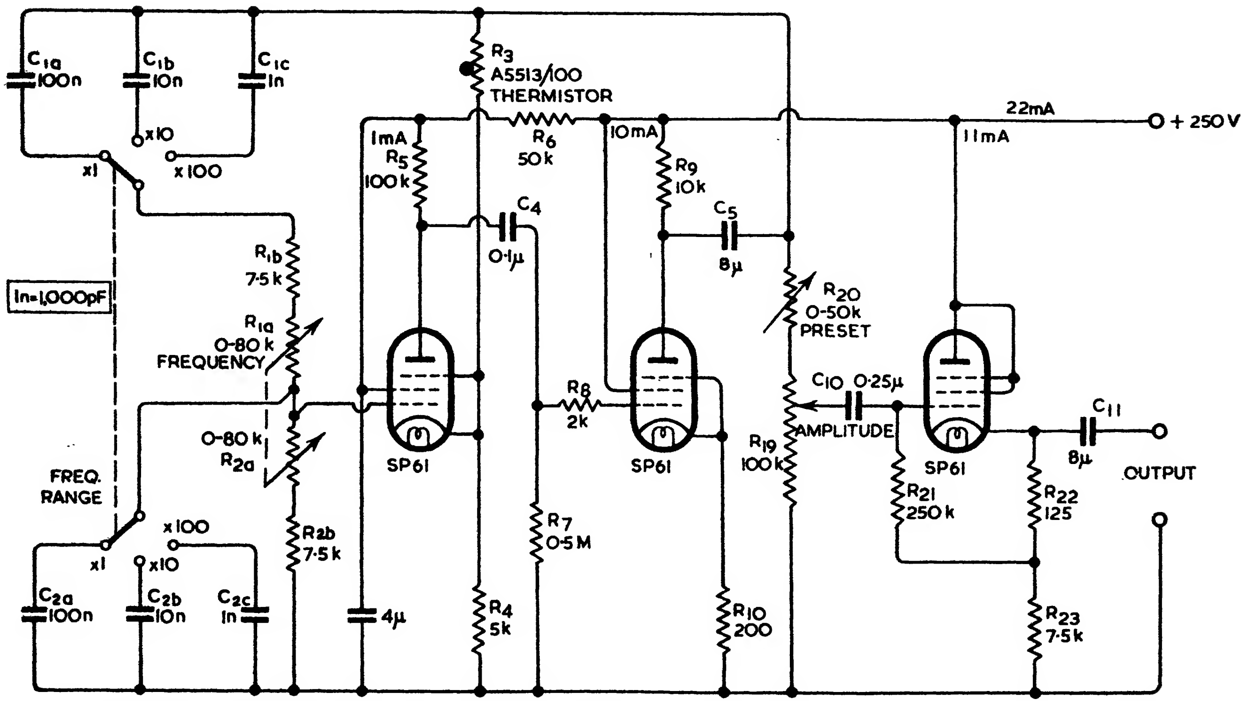


Fig. 4.34—Circuit diagram of 20 to 20,000 c/s audio source with amplitude stabilization and pure waveform

thermistor (R_8) keeps the amplitude constant within 0·2 dB over the whole range of frequency. R_{19} is used to control the input to the cathode follower (and thereby its output), and R_{20} prevents the input from exceeding an amplitude that can be handled with low distortion. The output into 5 kΩ is about 40 mW for a harmonic distortion of about 0·3%.

Calibration of such an oscillator is described in Secs. 10.10 and 10.11. It is convenient first to select C_{1b} and C_{2b} to give a 200 to 2,000 c/s scale with approximately equal margins above and below, and then calibrate fully over that range, afterwards adjusting the other capacitors to fit the same scale.

In general, calibration points do not by themselves form a convenient or adequate scale. It is advisable, therefore, to mark them temporarily in pencil on the dial, which is then detached and rolled along one axis of a sufficiently large piece of graph paper. As each mark comes against the paper, a corresponding mark is made on the axis. A frequency scale is then marked along the other axis, and a graph drawn connecting this scale with the marked frequencies. This graph can then be used to find the points on the dial corresponding to the wanted scale markings; they are transferred to it by another rolling operation.

A full description of a more elaborate generator, of which Fig. 4.34 is the nucleus, appeared (*W.W.*, Aug. and Sept. 1949). The additional features include a sine-to-square-wave converter, attenuators totalling 102 dB, a second output stage with phase inverter to give a balanced output, a monitor valve voltmeter of the type shown in Fig. 5.27 (but simpler, having only a single range), and conventional power unit. A square waveform is extremely useful for testing the transient response of amplifiers. A two-phase output, with one amplitude variable, is in many respects equivalent to a Wagner earth (Sec. 7.7) when the generator is used for a.f. bridges, but without most of the complications; and it increases the usefulness of the generator for many other purposes.

The SP61 valves could be replaced by modern equivalents such as EF80, and the frequency coverage extended by adding at least one range. This is virtually what was done by Sinfield in a very similar design among those listed below:

D. O. Roe and W. Morle, *S.R. and R.*, Aug. 1955. 20 c/s to 100 kc/s; Wien bridge; variable capacitance; thermistor stabilization.

L. F. Sinfield, *W.W.*, Dec. 1954. 6 c/s to 70 kc/s; Wien bridge; variable resistance; thermistor stabilization.

R. Williamson, *W.W.*, Oct. 1956. 3 c/s to 300 kc/s; Wien bridge; variable resistance; thermistor stabilization.

L. H. Dulberger and H. T. Sterling, July 1958. 0·9 c/s to 510 kc/s; Wien bridge; variable capacitance; lamp stabilization.

A. R. Bailey, *E.T.*, Feb. 1960. 10 c/s to 100 kc/s; high-gain amplifier with parallel-T feedback network giving distortion < 0·01% over most of the range and < 0·02% over all; variable resistance; thermistor stabilization.

W. D. Ryan and F. E. Hetherington, *E.E.*, Feb. 1960. 300 to 2,400 c/s; modified Wien bridge; voltage tuning; thermistor stabilization; 3–7% distortion.

F. Butler, *W.W.*, Aug. 1960 (see also Dec. 1960, p. 608 and Jan. 1961, p. 35). 20 c/s to 20 kc/s; transistorized Wien bridge; variable resistance; thermistor stabilization.

There are also "kits":

Daystrom Ltd., Heathkit AG-9U. 10 c/s to 100 kc/s; bridged-T; switched resistance; lamp stabilization.

Jason Motor & Electronic Co., AG.10. Sine wave, 10 c/s to 100 kc/s; square wave, 10 c/s to 50 kc/s; Wien bridge; variable capacitance; thermistor stabilization.

(The same firm supplies an a.f. attenuator as a kit or ready-made.)

4.29. R.F. SOURCES

A very large number of assorted radio-frequency oscillators are available absolutely free (apart from the cost of a receiving licence), and use of them for testing purposes fulfils one of the principles laid down in Chapter 3—the desirability of carrying out tests as nearly as possible under actual working conditions. The frequencies of certain B.B.C. stations are kept constant within a few parts in 10^7 (Sec. 14.37), which is better than most laboratory standards—and less likely to be influenced by stray coupling from the apparatus being tested! But although sending stations are very useful test oscillators, they have the disadvantage of not being under the control of any but exceptionally privileged workers. For most quantitative tests a signal modulated by a constantly fluctuating broadcast programme is unsuitable. And the signal strength from the more distant stations cannot be counted upon to remain absolutely constant. Even in simple tests on receivers one very soon experiences a need for a signal generator under one's own control.

From oscillators that can be assembled in a few minutes from spare parts, to standard-signal generators costing hundreds of pounds, there is a continuous range of available equipment. A rough division into three classes may usefully be made, however:

(*a*) Open oscillators, which may be calibrated in frequency, but in which no special provision is made for controlling the signal strength to repeatable levels.

(*b*) Oscillators, generally described as for servicemen's use, which are entirely screened except for a definite outlet to which the signal strength may be adjusted for comparative purposes.

(*c*) Standard-signal generators in which the signal control is calibrated in microvolts and in which other refinements, such as variable modulation depth, may be included.

At one time a heterodyne wavemeter, which was just a valve oscillator calibrated in wavelength, was the first—and sometimes almost the only—instrument in a radio laboratory; but nowadays a reasonably accurate frequency calibration is expected as one of the facilities of every signal generator. The accuracy of frequency measurement is so high that standards of frequency (which are dealt with in Sec. 6.13) have to be much more refined than the old wavemeters. An appropriate type of open oscillator is very useful for many kinds of r.f. measurement, but as its design is so much bound up with the measurement technique it is dealt with in Chapter 9 (Sec. 9.27). Attention can therefore now be concentrated on signal generators, especially standard-signal generators, for if their rather stringent requirements are understood there should be no difficulty with the simpler servicemen's signal generators, which incidentally are very useful even in the laboratory.

4.30. STANDARD-SIGNAL GENERATORS

For testing receiver design and some other important purposes a standard-signal generator is practically indispensable. Unfortunately a good one is expensive. In principle it is extremely simple—a valve oscillator variable over appropriate frequency ranges, and thoroughly screened to prevent uncontrolled radiation; a variable attenuator for controlling the signal output through the intended channel; and provision for modulating. The extent and refinement of control varies greatly in different models, and so, of course, does the price. As details of this sort can be obtained in abundance from makers' catalogues, the most profitable thing to consider here is what features are most useful in practice. Having done that, one is better able to select the most suitable generator.

4.31. NECESSITY FOR THOROUGH SCREENING

The range of output in a laboratory generator is desirably from 1 μV to 1 V at least. The lower figure, or even less, is necessary for measuring the more sensitive types of receiver; the higher is needed for discovering spurious responses under extreme local station conditions, or taking a.g.c. curves. An oscillator powerful enough to give an output of 1 V in a circuit which prevents the external circuit reacting back to the oscillator must be very completely screened if the leakage field is to be imperceptible when working at the microvolt level. That is the first problem, and it involves much more than simply enclosing the apparatus in a metal-lined box. A very small current passing, say, between two earthing connections on the screen might be enough to set up an appreciable external field. In mains-driven models the mains lead must be very thoroughly filtered to prevent escape of r.f. there. When trying a signal generator it should be connected to a sensitive receiver at the highest signal frequency in the range, and the attenuator setting gradually reduced from a few microvolts to zero. If the receiver is sensitive enough to give an output which is substantial on one microvolt, but which steadily decreases to nil at the zero setting, then the generator screening may be considered satisfactory.

4.32. THE ATTENUATOR

Next, there is the design of attenuator to deal with such a large range of signal strength—120 dB—at high frequencies. Attenuators in general are dealt with in Secs. 6.18 to 6.22. The problems of their

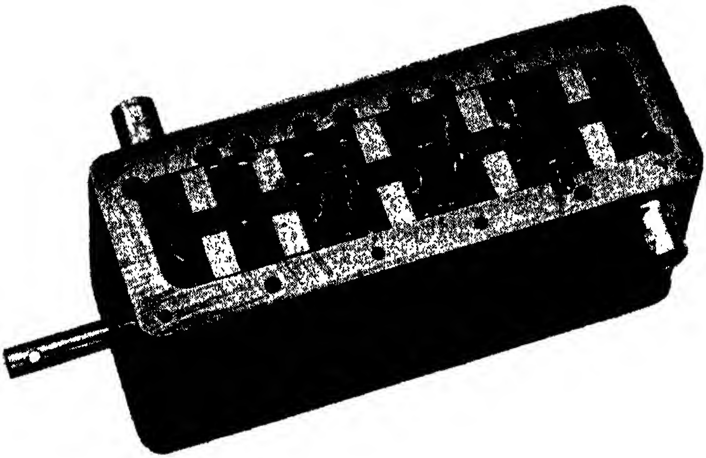


Fig. 4.35—Internal construction of Hatfield decade attenuators Type Q, for use in the frequency range 0 to 500 Mc/s. They are available for 50Ω or 75Ω, with 0 to 11 dB in 1dB steps or 0 to 110 dB in 10 dB steps. (Hatfield Instruments, Ltd.)

design are particularly acute in signal generators, where such a large range of attenuation has to be provided over such a large range of frequency (*M.I.*, Sept. 1954, pp. 150–9). The higher the frequency the more difficult it is to prevent the attenuator being by-passed by stray capacitance or rendered inaccurate by series inductance. The problem is sometimes eased in the simpler instruments by providing variable output over a limited range, say 1 to 10,000 μV, supplemented by a " force " output of perhaps 1 V. The most usual type of attenuator is the ladder (Sec. 6.21) varied in steps of 10 or 20 dB, in conjunction with a continuously variable slider, or variation of the oscillator amplitude. Wire-wound resistors have to be very carefully designed to be reliable up to 30 Mc/s. Fig. 4.35 shows the construction of a 110-dB resistive ladder attenuator, variable in steps of 10 dB, and useful up to 500 Mc/s. The elements are high-stability carbon resistors in an aluminium die-casting shaped to give good screening and to match 500 or 75 Ω impedance.

The same design is obtainable with steps of 1 dB; and one of each in tandem gives 121 dB in steps of 1 dB, which is adequate for most signal generators at all frequencies from zero up to at least 300 Mc/s. For still higher frequencies it may be necessary to use a piston attenuator (Sec. 6.22). Where attenuation is variable only in steps, the input should be variable continuously over at least the range of a step.

The accuracy of a microvolt calibration, which may be within 5-10% at medium frequencies, often deteriorates very considerably at the highest frequencies. Other sources of error in such work would make higher accuracy of little significance even if it could be obtained. When examining a generator, with a receiver and output meter connected so as to indicate equality of signal, some clue to attenuator accuracy can be gathered by noting whether the maximum signal at one setting of the multiplier switch is the same at one-tenth on the next higher " ×10 " switch position. Another is to measure the maximum output with a valve voltmeter. But most of the accuracy must be left to the maker's reputation, or to a N.P.L. calibration, unless one has suitable measuring equipment. One method* is to use the superheterodyne principle to convert the generator output to some relatively low frequency, at which the calibration of an attenuator is dependable.

To eliminate errors due to changes in the output from the oscillator, which is bound to vary to some extent as the frequency is changed,

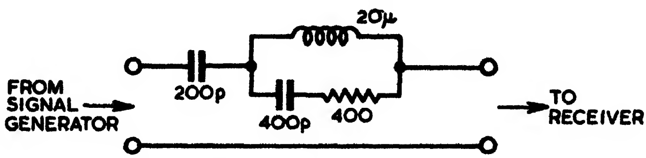


Fig. 4.36—Composition of standard dummy aerial

the input to the attenuator is monitored by a built-in meter. A thermal type has the advantages of reading true r.m.s. values at every frequency from zero up to the highest, but is easily burnt out. The thermionic diode has been generally superseded by the crystal diode, which has no " zero " current to balance out. A single diode measures peak voltage, which is not a good indication of fundamental r.m.s. if there is much even-harmonic content; in that case a voltage doubler (peak-to-peak) is more reliable (Sec. 5.14).

To ensure that the signal level at the point of test is the same as at the outlet of the generator, the link between must be a reasonably loss-free cable having a characteristic impedance equal to the generator output resistance. This matter, including the effects of incorrect cable impedance, is elucidated in detail by T. P. Flanagan (*M.I.*, March 1954). The correct calibration condition is with a resistive load equal to the nominal source impedance. The error resulting from the actual source impedance varying from this nominal value is then comparatively small; see D. Woods, *Proc. I.E.E.*, Part B, Jan. 1961, p. 38.

4.33. THE DUMMY AERIAL

The aerial-earth impedance of a receiver is not infinite, and the voltage established between those terminals depends on the impedance in series with them externally. To standardize generator measurements on receivers, a compact dummy aerial, made up as in Fig. 4.36, is inserted between the generator and the receiver under test. This

* " A Method of Calibrating Standard-signal Generators and R.F. Attenuators ", by G. F. Gainsborough. *J.I.E.E.*, Part III, May 1947.

composite impedance, shown in Fig. 11.23, has a minimum value of 220 Ω (at about 2 Mc/s). For precise comparisons the generator output resistance at the dummy aerial input terminals should be some standard figure, say 75 Ω, and any deficiency made up by resistance incorporated in the dummy aerial or inserted between it and the generator. In some signal generators the output resistance is considerably greater at maximum signal voltage than on the lower ranges; if so, this may lead to false comparisons when a low-impedance load is being fed, unless compensation is made by external resistance.

4.34. FREQUENCY CONTROL

Ideally, the attenuator setting and the nature of the circuit connected to the outlet ought to have no effect whatever on the frequency, but in practice this is not always so, particularly at the highest frequencies and in generators with no buffer stage between oscillator and attenuator; and unless the generator tuning is checked after altering the attenuator setting, results may mislead. It is quite easy to discover this defect by trial; and it is most serious when making selectivity measurements.

The inclusion of a buffer stage and other refinements adds to the range-switching problem, which is worst at very high frequencies. Some of the best instruments use turret-mounted coils, to retain the same active circuit layout on all frequency bands and to minimize connecting leads. In some models the problem of switch contacts at v.h.f. is avoided by using capacitance switching.

The tuning capacitor ought to allow close adjustment, and a subsidiary control for varying the frequency slightly above and below that of the main dial is a valuable feature for selectivity tests. Direct-reading frequency scales are very much more convenient to use than calibration curves, provided that the accuracy is reliable. As a signal generator is expected to do duty as a frequency meter, the accuracy and stability of calibration is important.

4.35. MODULATION

The cheapest instruments are generally modulated to an uncertain depth; in the better sorts the modulation depth is not only accurately known but is variable at will. The depth usually assumed for general use is 30%, but for some purposes it is necessary to be able to vary it over as wide a range as possible, preferably up to or very near 100%. For the same purposes—tests of detector distortion, etc.—the distortion due to the generator, even at maximum depth, ought to be, but seldom is, negligible. Depth and linearity of modulation can be checked by cathode-ray oscilloscope (Sec. 11.23).

Another thing to be avoided in amplitude modulation is frequency modulation. Here again a buffer stage is the chief safeguard.

Amplitude modulation is quite easy if the foregoing defects are tolerated, but not if a high standard is required. In avoiding f.m. by modulating an amplifier stage instead of the oscillator, one is likely to distort the carrier wave severely, and if this is to be corrected it is necessary to tune the output. Suppressor-grid modulation is quite good, especially if not required to exceed about 80%. Anode and even screen-grid modulation require appreciable power. Control-grid modulation is perhaps the most likely to give trouble. The Marconi Instruments TF 1102 Amplitude Modulator,* which can be attached to any signal source, consists basically of a common-grid valve.

The usual fixed modulation frequency for a.f. work is 400 or 1,000 c/s, and for most purposes the close accuracy of this is immaterial. Provision is generally made for external modulation, for taking a.f. characteristics; it is an advantage if too much power is not required, especially at the upper frequencies. Some generators can be modulated straight from a gramophone pick-up, others require a large fraction of a watt.

A useful feature, particularly in the cheaper models, is a switch for enabling the modulation-frequency signal to be used externally. And a switch for giving an unmodulated radio signal is definitely necessary.

With increasing use of pulse and frequency modulation for communications, signal generators with these facilities are becoming available. A convenient method for f.m. is to superimpose modulating current on a fixed current through a saturating coil wound around the ferrite core of a coil forming part of the r.f. tuning circuit. Other methods are described in the next Section.

In signal generators for television, modulation is the main issue, and a set of modulation waveforms should be obtainable for testing scanner linearity, definition, synchronization, etc. To make a complete check of synchronization and interlacing, the modulation must conform accurately to broadcasting standards, and this calls for quite an elaborate instrument. A pattern generator using 86 transistors and 73 crystal diodes, conforming to B.B.C. standards, has been described by J. Schaffer and D. W. Furby (*Proc. I.E.E.*, Vol. 106, Part B, Supp. 18, May 1959, p. 1260).

4.36. "WOBBULATION"

This is a specialized type of frequency modulation, used in conjunction with an oscilloscope for observing frequency responses (Sec. 5.38). The signal generator frequency is swept to and fro at a low rate—50 c/s at most—over a selected band. An obvious method is to rotate an auxiliary tuning capacitor by a motor. Although this is not generally favoured now, the author has found it successful for laboratory work, and it does allow any desired law of frequency variation to be arranged by suitably shaping the capacitor vanes. A gramophone type of motor can be used to rotate them at about 300 r.p.m. One revolution gives two frequency sweeps each way,

* "A.M. without F.M.", by N. G. Webb. *M.I.*, March 1956.

and a contact mounted on the spindle is used to synchronize a linear time-base circuit (Sec. 5.33). If the capacitor fixed vanes are arranged in several sections the capacitance sweep can be varied in steps according to the number in circuit. If it is not practicable to enclose this gear in the signal generator, it can be mounted in a screened box, with screened lead for connecting in parallel with the generator tuning circuit.

An alternative mechanical method is to use a loudspeaker type of moving coil as the drive for a coaxial capacitor. It is conveniently energized from the 50-c/s supply. An advantage is that the frequency

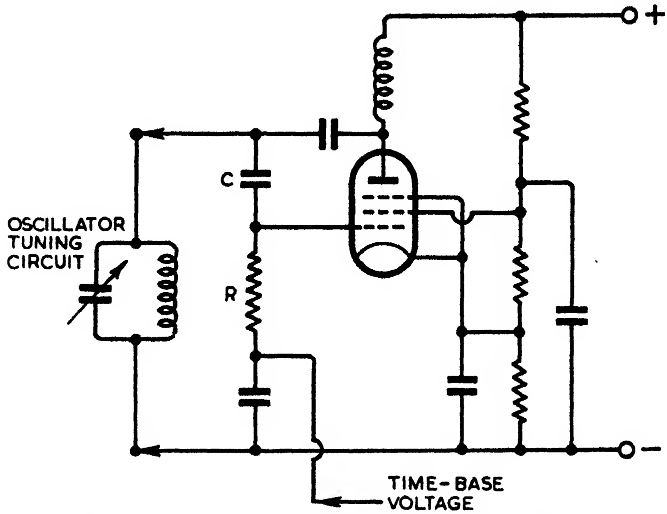


Fig. 4.37—Reactance-valve circuit, for causing the voltage of a time-base to produce corresponding frequency variations in the signal source

sweep can be continuously varied down to zero by controlling the moving-coil current. A simple practical example is described by B. T. Gilling (*W.W.*, June 1956).

Most often, however, an electronic method is used. One basis for this is a valve made to serve as a reactance by feeding its control grid with the signal voltage through a phase-shifting circuit, such as C and R in Fig. 4.37. Provided that the reactance of C is much greater than R, the voltage at the grid leads the signal voltage by nearly 90° and the resulting anode current is in the same phase as if the valve were a capacitor across the signal source. The amount of the current, and hence the apparent capacitance, depends on the mutual conductance of the valve; so the frequency can be varied by the time-base voltage as shown.

A similar result is obtained in a different way due to K. C. Johnson, and described by him (*W.W.*, April and May 1949). Advantages are that as much as 30% frequency modulation is possible, and the oscillator valve itself can be made to do its own modulating. The author's design for a single-valve Johnson wobbulator with a maximum sweep of 180 kc/s at about 1 Mc/s (*W.W.*, Oct. 1950) is obtainable from the publisher in leaflet form, together with a simple oscilloscope for use with it. The circuit diagram is shown here as Fig. 4.38.

Another method is by ferrite core, referred to in the previous Section. Yet another uses a semiconductor diode as a voltage-controlling tuning

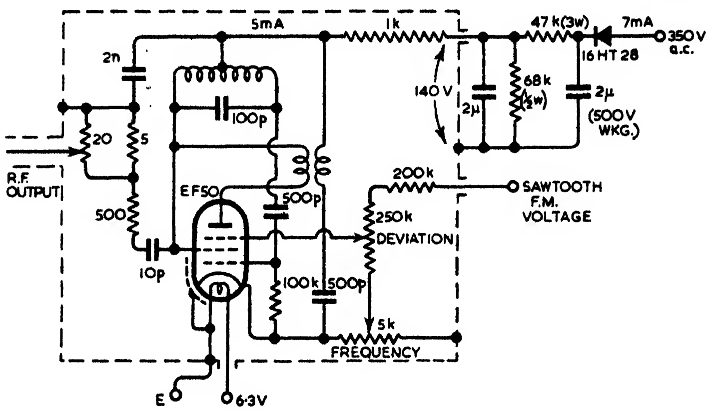


Fig. 4.38—Author's practical version of Johnson one-valve f.m. oscillator, or "wobbulator"

capacitor, as described by G. G. Johnstone (*W.W.*, Aug. 1956).

RC oscillators can conveniently be frequency-modulated by replacing resistance elements with valves or transistors. Details of this scheme are given by G. W. Short (*W.W.*, Nov. 1959).

A f.m. signal generator with provision for external modulation usually forms a good nucleus for a wobbulator system.

To keep the frequency sweep of a wobbulator the same when the tuning of the oscillator is varied, the beat-frequency technique can be used (Sec. 4.26), wobbulation being applied to what would otherwise be the fixed-frequency oscillator.

There is an excellent general discussion of wobbulators by R. Brown (*W.W.*, Jan. and Feb. 1961); see also *B.C. & E.*, April 1960.

4.37. COMMERCIAL R.F. SIGNAL GENERATORS

The comparatively inexpensive signal generators designed primarily for broadcast-receiver servicing, and making no claim to precision of output level, are nevertheless very useful in the laboratory where funds are limited. A good example in this class is the Avo Type III, covering 150 kc/s to 220 Mc/s in six ranges. Output is from about 1μV to 100 mV continuously variable, with an additional fixed level of 0·25 V.

The signal can be either unmodulated or amplitude modulated. Fig. 4.39 shows the interior.

A rather more elaborate signal source by the same maker, Type TFM, provides both a.m. and f.m., and includes a device for adjusting the frequency scale so as to bring its markings precisely into line with an accurately known signal.

An inexpensive instrument covering the exceptionally wide range 150 kc/s to 390 Mc/s is the Cossor Model 1450.

One of the cheapest is the Channel Electronic Industries Type 40—a miniature a.f. and r.f. transistor oscillator covering up to 20 Mc/s on fundamentals.

An example of the laboratory class of signal generator, providing a wide range of facilities with specified accuracy, is the Airmec Type

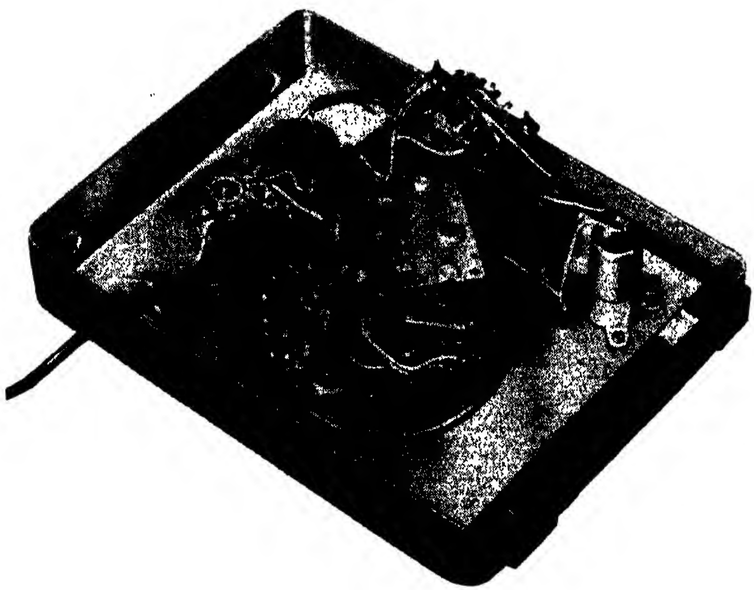


Fig. 4.39—Interior of AVO Type III r.f. signal generator. The whole apparatus is screened by the outer metal case; the oscillator circuits are doubly screened by the inner cylindrical cover which has been removed. (AVO, Ltd.)

204. The frequency range is 1 to 320 Mc/s, and provision is made for c.w., a.m., f.m., external p.m. (pulse modulation), or combined a.m. and f.m. The 1 to 20 Mc/s band is obtained by combining the output from two higher-frequency oscillators, and the remaining four bands by successive frequency-doubling from a 5 to 10 Mc/s oscillator. A built-in crystal calibrator provides frequency check points on all ranges, and f.m. deviation check.

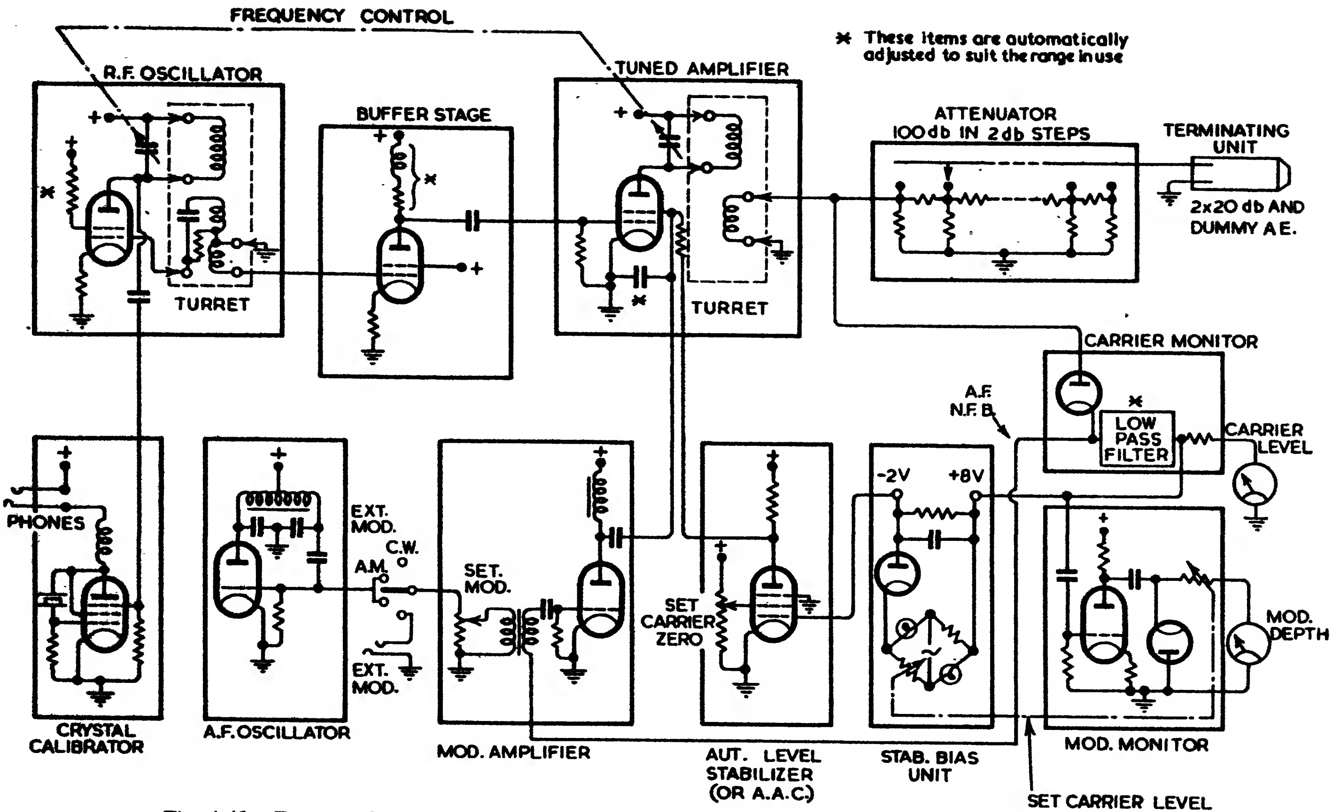


Fig. 4.40—Functional diagram of a laboratory standard-signal generator—Marconi Instruments TF.867

Fig. 4.41 — Oscillogram of modulation envelope with modulating waveform superimposed, showing fidelity of modulation even at practically 100% depth. (Marconi Instruments, Ltd.)

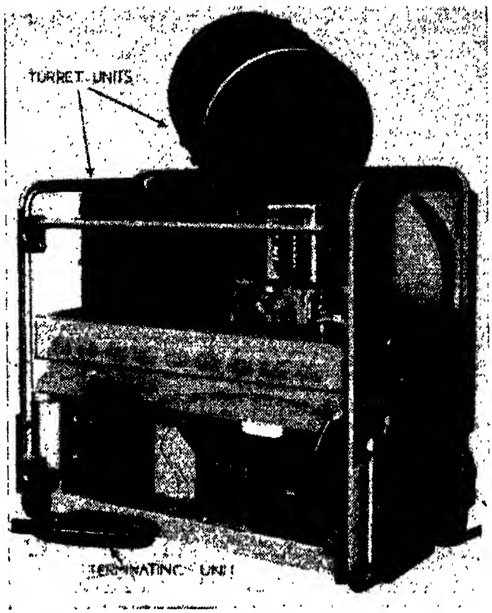


Fig. 4.42 — Opened-up rear view of TF.867 standard-signal generator, with one coil turret lifted off. (Marconi Instruments, Ltd.)

Of generators that provide a modulation conforming to British television broadcasting standards, enabling all essential receiver tests to be carried out, the Telequipment WG/44 is notable for small size and comparatively low cost. It provides a choice of five patterns on a carrier wave variable from 40 to 70 and 170 to 220 Mc/s.

What can properly be described as *standard* signal generators are usually designed for frequencies either lower or higher than about 30 Mc/s, but not both, as the techniques differ. A particularly interesting example covering the lower frequencies, from 30 Mc/s right down to 15 kc/s—and so including most of the v.f. as well as communication and sound broadcasting bands—is the Marconi TF.867. Fig. 4.40 is a skeleton diagram showing its functions. Note that the oscillator is separated from the attenuator by two valves. The modulation frequency operates on the second of these through a modulation valve, and up to 100% modulation can be obtained without frequency modulation. Distortion of the modulation envelope is removed by rectifying the modulated carrier wave and using the m.f. output for negative feedback to the modulator; the oscillogram Fig. 4.41 shows the result. Depth of modulation is measured by rectifying the actual carrier wave—a more reliable method than deducing it from the a.f. modulating voltage. Carrier amplitude is kept constant by a.a.c., and its frequency can be checked by a crystal calibrator. In such an elaborate system the range switch is required to change over a large number of contacts, and Fig. 4.42 shows the double-turret system.

Output is variable over the exceptionally wide range of 0·4 μV to 4 V (140 dB) in fifty 2-dB steps by means of the attenuator shown in Fig. 4.43, supplemented by two steps of 20 dB in the terminating unit.

Aspects of standard-signal generator design are reviewed from time to time in *Marconi Instrumentation* (*M.I.*), obtainable from Marconi Instruments, Ltd. Details of types for testing mobile receivers (a.m. and f.m.) are given by J. F. Golding (*W.W.*, Feb. 1960). Three portable transistor a.m. signal generators, also mainly for mobile receivers, are available from R.E.E. Telecommunications Ltd. Model A covers 40 to 70 Mc/s; Model B, 100 to 150 Mc/s; and Model C, 70 to 72 and 85 to 87 Mc/s. The design of a transistor signal generator for 100 kc/s

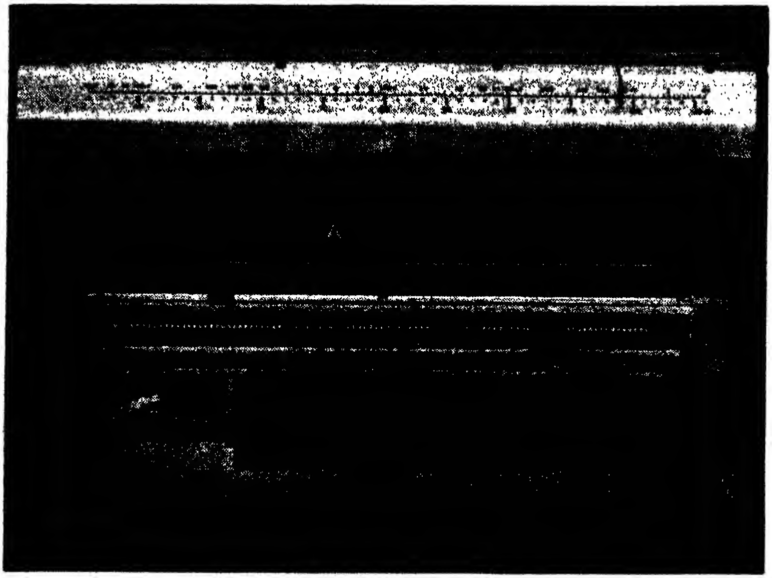


Fig. 4.43—Front (a) and back (b) of 100-dB attenuator used in TF.867 standard-signal generator. An additional 40 dB is obtainable with the terminating unit seen at the foot of Fig. 4.30. (Marconi Instruments, Ltd.)

to 25 Mc/s is detailed by C. Bayley (*W.W.*, Jan. 1961). Commercial standard signal generators have been reviewed by R. Brown (*B.C. & E.*, Jan. 1961.).

4.38. SPECIAL WAVEFORM GENERATORS

The basic form of signal is the sine wave, but developments such as television and radar have stimulated the production of other waveforms, which are being adapted increasingly to laboratory uses; in bridges, for example. The most important forms* are the square

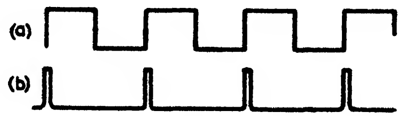
* The sawtooth waveform used for time bases is considered in connection with oscilloscopes in Sec. 5.33.

and pulse, shown as *a* and *b* respectively in Fig. 4.44. The difference between them is of degree rather than kind—the relative duration of the positive and negative phases of the cycle, or the mark/space ratio, as it is more briefly termed. When this is 1 : 1, or not very far off, it is a square wave; when the ratio is large it is a pulse.

There are two main classes of methods: in one, an existing waveform (usually sinusoidal) is shaped into the desired form by distorting devices; in the other, the waveform is generated directly, though the times of its occurrence may be controlled by applied signals. Details of both are given in almost any book on radar.

If what is required is a square wave, and a sine wave generator is available, the first policy is simple, economical and convenient. In

Fig. 4.44—Square-wave (*a*) *and pulse* (*b*) *waveforms*



particular, no frequency-controlling devices are needed. One common method is to pass the sine wave—at as large an amplitude as possible —to a valve with a very short grid base to cut off the negative peak by bottom bend, and a high resistance in series with the grid to cut off the positive peak by grid current. More than one "squaring" stage may be needed to produce a steep enough rise and fall. Details of a self-contained squarer for adding to an a.f. oscillator is given by A. W. Wayne (*W.W.*, Feb. 1955), and a miniature transistorized squarer deriving its modest power supply as well as its signal from the sine-wave output of the generator to which it is attached is described by J. C. S. Richards (*E.E.*, Nov. 1960, p. 720).

Unless special provision is made for biasing the input, the foregoing method is likely to give unequal positive and negative "half" cycles. A more symmetrical arrangement is the "long-tailed pair" (Fig. 4.45) using high-μ valves and low anode voltage. The fact that two outputs of opposite polarity are available may be useful. And it can be designed for frequencies up to the megacycle order.

The resulting approximation to a square wave can be turned into a pulse of controllable width by means of a differentiating circuit—a capacitor and resistor in series having a time constant much less than the period of one cycle; the output is taken from across the resistor, and if necessary can be squared off by a limiter. The shaping technique is simple and straightforward, but a large amount of amplification is needed to make the wavefronts really steep.

The self-generating types are usually *RC* oscillators of some kind, and may either be self-repeating (but capable of being "locked" to an alternating signal within certain limits of frequency) or "one-shot" devices which have to be triggered by an applied pulse. Most modern pulse generators consist of both; one to control the recurrence frequency and the other to control the width (i.e., duration) of each pulse.

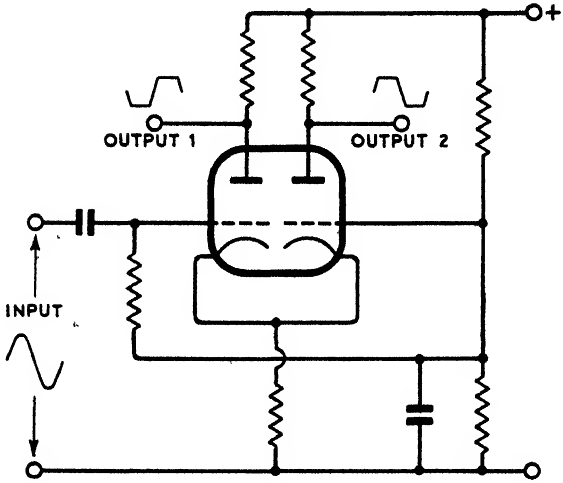


Fig. 4.45—" Long-tailed pair " squarer, which receives a large-amplitude sine wave and supplies two approximately square waves of opposite phase

A transitron with a resistance in place of the tuned circuit can be used to generate very good square waves, with mark/space ratio controllable in both directions; and a practical design for three fixed frequencies—80, 800 and 8,000 c/s—for such purposes as a.f. testing in conjunction with an oscilloscope is given by O. C. Wells (*W.W.*, Jan. 1951). The classical circuit for producing steep-fronted (though not usually flat-topped) oscillations is the multivibrator. In its conventional form it consists of two *RC* amplifier stages, each feeding the other, basically as in Fig. 4.46.

A simpler and better multivibrator circuit, giving a much squarer waveform, is a variety of the cathode-coupled oscillator described by K. A. Pullen (Sec. 4.22) and is derivable from Fig. 4.28 by shorting out R_1 and omitting C_2. It is also known as the Schmitt circuit. A simple and inexpensive square-wave generator based on it, variable from 15 c/s to 160 kc/s, is described by L. F. Sinfield (*W.W.*, July 1952). The full circuit is shown here as Fig. 4.47. With care in minimizing stray capacitance, particularly to the capacitors, the time of rise or fall of the waveform can be made less than 1 μsec.

Designing for pulses of specified width is dealt with in:
" Cathode-Coupled Flip-Flop " by T. G. Clark, *W.W.*, Jan. 1958.
" The Schmitt Multivibrator ", by G. L. Swaffield, *W.W.*, July 1958.

Transistor multivibrators are similar to Fig. 4.46 except that as usual the " base leaks " are returned to the collector side of the power supply. Design and performance are discussed by A. E. Jackets (*E.E.*, May 1956). F. Rozner has shown (*E. & R.E.*, Jan. 1957) that

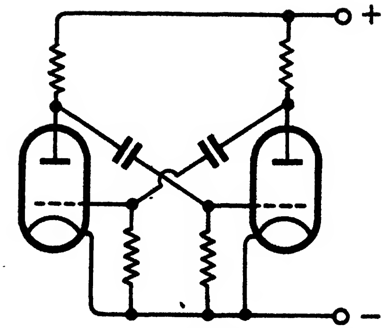


Fig. 4.46—Basic multivibrator circuit

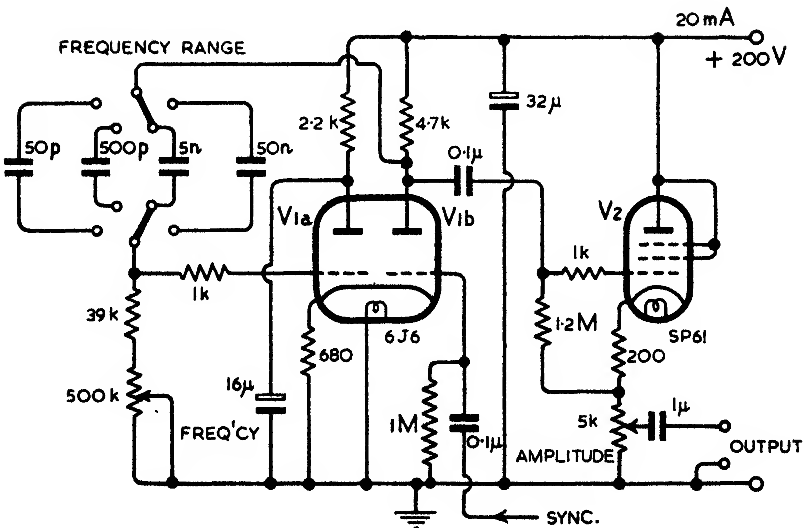


Fig. 4.47—Circuit diagram of 15 c/s to 160 kc/s square-wave generator

pnp and *npn* transistors can be paired to give pulses of 0·7 μsec at 100 kc/s with a peak power of 1 W. A variety of transistor multivibrator especially suitable for frequency division is included in Sec. 6.15.

For rapid testing of television lines and amplifiers, a signal form known as " pulse and bar " has been standardized.* It consists of a 0·17 μsec or 0·33 μsec sine-squared pulse, a 40 μsec rectangular pulse and a 10 μsec negative pulse, as in Fig. 11.19. A commercially available generator of this composite waveform is the Pye PTC 1201/3.

For general pulse work with oscilloscope the need is for a versatile generator, providing a wide range of frequency, amplitude and pulse width, and perhaps a " pre-pulse " leading to the main pulse with variable delay. The Rank Cintel pulse and sweep generator has all this, plus a sawtooth waveform.

Current commercial practice in pulse and square-wave generators was reviewed in *E. & R.E.*, June 1959.

One of the properties of a sharp-cornered waveform is its large number of harmonics—theoretically an infinite series of them in a perfectly rectangular wave. This wide frequency spectrum of signals is sometimes the object in view; for example, in the multivibrator, used chiefly for extending a single accurately-known frequency to a large number of points (Sec. 6.15). For this purpose the precise waveform is unimportant so long as the front is extremely steep and the frequency can easily be locked to that of another signal. A convenient place for feeding a synchronizing signal into the circuit shown in Fig. 4.47 is the grid of V_{1b}.

The theory and practice of multivibrators is given in some detail in E. K. Sandeman's *Radio Engineering* (Chapman & Hall, 1947) Vol. 1, XII: 6.2.

Another multivibrator application is as a generator that virtually gives a signal at all frequencies simultaneously. In one (*W.W.*, April 1939), the fundamental frequency is 400 c/s, and harmonics up to and beyond the 50,000th (20 Mc/s) are detectable. This is particularly useful for adjusting the oscillator padding of superhets. The circuit given in Fig. 4.47, perhaps with slight modification to equalize the harmonics, would be suitable. The same principle is employed in " pen-torch " signal generators, which are really buzzers generating a wide-spectrum signal for receiver checking.

4.39. " NOISE " GENERATORS

The type of signal just mentioned, on thousands of closely-spaced frequencies at once, is half-way from the pure single-frequency signal to its opposite extreme—one occupying all frequencies with no spaces between. Being the electrical analogue of white light, it is often called " white noise ". It is generated unavoidably in every circuit and valve by the random movements of electrons, and is the ultimate limiting

* " Waveform Responses of Television Links ", by H. W. Lewis, *Proc. I.E.E.* Part III, July 1954.

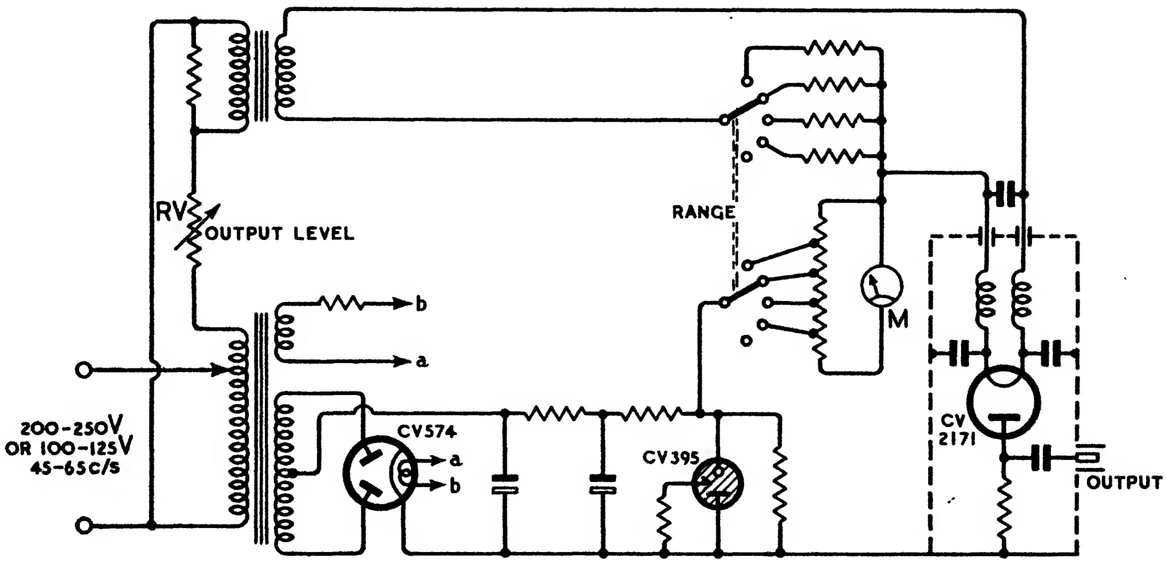


Fig. 4.48—Circuit diagram of Marconi Instruments TF.1106 " noise " generator

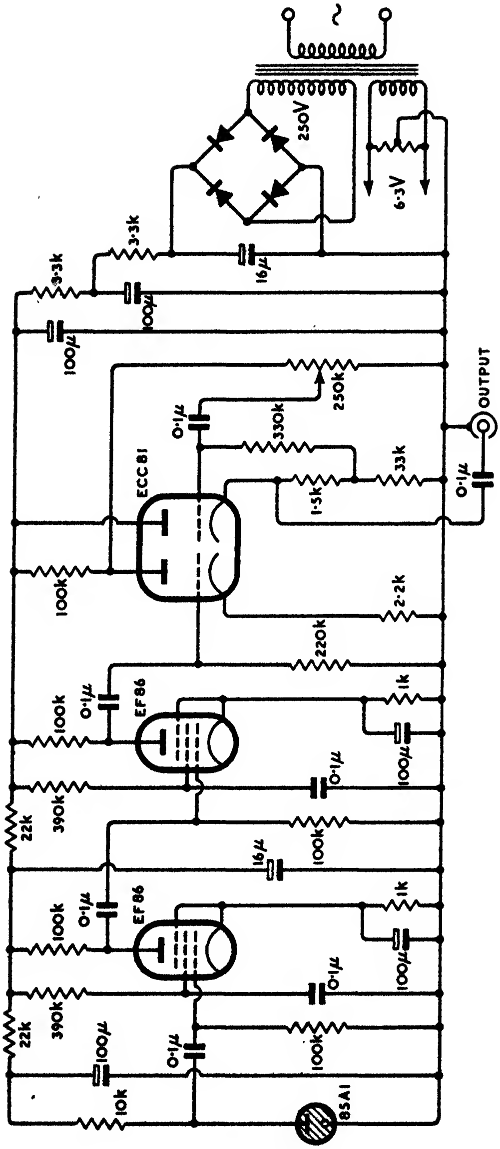


Fig. 4.49—Circuit diagram of noise generator using a diode followed by a high-gain aperiodic amplifier

factor in detecting weak signals, for the greater the amplification the greater the output of noise.

The noise power produced in a circuit or valve can be calculated (Sec. 14.32), and this is used to provide a standard low-level signal, particularly at very high frequencies where it is difficult to do this with a conventional signal generator. In particular, a noise generator is a convenient instrument for measuring the most important figure of a receiver for v.h.f. or over—its *noise factor*.

A suitable standard-noise generator is a temperature-limited diode; i.e., one in which the flow of current is limited only by the temperature of the cathode and not by the anode voltage. This is to prevent the formation of a space charge, which would reduce the noise level and make it much more difficult to calculate. The Marconi Instruments TF.1106 embodies this principle; Fig. 4.48 is its circuit diagram. The noise-factor range switch controls both the filament voltage (and therefore its temperature) and the anode milliammeter shunt. The noise output is adjusted by the continuously-variable filament control RV, and the anode current (to which the noise power is proportional) is indicated by M. The noise output is taken from the anode via a 52-Ω or 71-Ω termination to which the receiving system under test is matched.

The TF.1106 is for r.f. tests (actually 1 to 200 Mc/s), but white noise is of great value in a.f. testing, as it can be generated cheaply and saves thousands of pounds that would otherwise have to be spent on an anechoic chamber. Fig. 4.49 is a circuit diagram given by J. Moir (*S.R. & R.*, Feb. 1959) of a noise generator due to J. Newton. It consists of a 85A1 stabilizer tube as the noise source, followed by a high-gain amplifier. To avoid hum, care is required in the layout of the first stage. Its heater and grid leads should be screened, and the mains transformer placed as far away as possible. A piece of metal tube packed with cotton wool may have to be placed over the 85A1 to reduce acoustic feedback.

For carrier frequencies, Marconi Instruments have a 12 to 4188 kc/s white noise generator—TF1226B. (*M.I.*, March 1961).

References:

"Theory and Measurement of Noise Factor", by R. J. Yates. *M.I.*, July–Aug. and Oct.–Nov. 1950.

"A Generator of Electrical Noise", by A. P. G. Peterson. *G.R.E.*, Dec. 1951.

"Physical Sources of Noise", by J. R. Pierce. *Proc. I.R.E.*, May 1956.

"Methods of Solving Noise Problems", by W. R. Bennett. *Proc. I.R.E.*, May 1956.

Noise, by Van der Ziel (Chapman & Hall, 1955).

CHAPTER 5

Indicators

Having applied some sort of signal to the apparatus under test, one requires an indicating instrument to give a reading of the results. The most important classes of instruments are (1) " meters ", i.e. pointer instruments, without the aid of amplification; (2) valve and transistor voltmeters; and (3) cathode-ray equipment. These will now be considered in turn.

5.1. BASIC TYPES OF METER

In the great majority of instruments for measuring current, the force needed to deflect the pointer is derived from the interaction of two magnetic fields, one or both being due to the current being measured. The logical way of measuring voltage is to make use of the force of electric fields, and this is done in electrostatic voltmeters; but unfortunately the counterpart of a permanent magnet is lacking, so in practice this method is restricted to fairly high voltages, and voltage is more often indicated by the current it drives through a resistance. In thermocouple meters the current used to move the pointer is derived even more indirectly, from the heating of a metal junction by the current being measured.

Meters using magnetically-produced force can be divided into three main types: the permanent-magnet moving-coil (hereafter called just " moving-coil "), in which one of the fields is produced by a fixed permanent magnet and the other by a coil moving between its poles; the moving-iron, in which two pieces of iron, one fixed and one moving, are magnetized by a fixed coil; and the so-called electrodynamic or dynamometer, in which both fields are produced by current-carrying coils, one fixed and one moving.

5.2. CHARACTERISTICS OF METER TYPES

Current meters are usually designed so that the deflection is proportional to the force, and the force is proportional to the product of the two magnetic fields, and fields produced by coils are proportional to the current flowing. So in moving-coil meters, where one magnetic field is constant, the deflection is proportional to the current, which means that the scale is linear and therefore most easy to read. If the current varies too rapidly for the pointer to follow, the deflection is proportional to average current; and as the average value of each cycle of a.c. is zero (the positive and negative halves cancelling one another) the moving-coil type indicates only d.c. In the other two electromagnetic types, both fields are due to the current being measured, so the deflection is proportional to its square, giving a square-law scale, cramped near the zero. If the current being measured is varying rapidly, the r.m.s. value is indicated, so such meters can be used for either d.c. or a.c. The same applies to electrostatic voltmeters; and also to thermal meters, because heat is proportional to power, which is proportional to current-squared. Thermal meters have the particular advantage that the heat-developing resistance can easily be arranged to be reasonably constant at all frequencies from zero to hundreds of Mc/s, so they cover a wider range of frequency than any other. Unfortunately they are quickly burnt out by a slight momentary overload, and are either very expensive or power-consuming or both. The d.c. generated by the heat is read on a sensitive moving-coil instrument, suitably scaled. The heater and junction can be one and the same, or touching, or spaced apart slightly. It is desirable for them to be enclosed in a vacuum to exclude air cooling.

Ideally, a current meter should have zero resistance, so that it subtracts no voltage from the circuit. And a voltmeter should have infinite resistance, so that it subtracts no current. The actual voltage drop across a current meter and the current taken by a voltmeter (in both cases at full-scale deflection) should always be ascertained and, if significant, allowed for in measurements. The voltmeter current is often specified—for no clear reason—in the indirect form of ohms per full-scale volt. Thus, if the maker says 1,000 Ω/V, one can deduce that the current required to produce full deflection is 1 mA. Why not say so?

The great majority of meters are of the moving-coil type, because they can be adapted for measuring a very wide range of current and voltage with good accuracy and little consumption of power, and in cheapness are surpassed only by meters that are very little good in either of these respects. These qualities are due mainly to the fact that one of the magnetic fields is very strong all the time, whereas in the moving-iron and dynamometer types both the fields weaken as the current is reduced, and it is difficult to get a deflection without drawing an excessive amount of power. A linear scale is reckoned as being effective over a range of 1 : 10, whereas a square-law scale is effective over only about 1 : 3, so something like twice as many ranges are needed, though admittedly this disadvantage is sometimes reduced by specially shaping the movement to give a nearly linear scale over all except the low end.

Although the moving-iron and electrodynamic types have the advantage of reading either d.c. or a.c., the inductance of the coils, and their power consumption, limit their usefulness mainly to power-supply readings. At one time moving-iron meters were noted for inaccuracy, and this still holds for cheap specimens, but they can now be obtained with accuracy comparable to moving-coil if one pays handsomely for it. It should be noted that even then the d.c. readings are likely to differ appreciably according to direction of current, and their mean may not be exactly the same as the a.c. reading. The electrodynamic type is similarly limited in frequency, but can be made very accurate, so is used as a standard for calibrating other a.c. meters. And having two coils, one (current coil) can be designed for connecting in series with a load and the other (voltage coil) in parallel, the reading then being proportional to the power taken; in other words, an electrodynamic instrument can be designed as a wattmeter (Sec. 5.6).

An unfortunate thing about the moving-coil meter is that there is nothing quite so good all-round for a.c. The result is that in most popular general-purpose meters a rectifier is incorporated to turn a.c. into d.c. so that it can be measured on the moving-coil meter used for d.c. measurements. Except at low voltages, the a.v. scale is practically linear, and the deflection is proportional to the average value of the half-cycles. With a sine waveform this is $2\sqrt{2}/\pi = 0{\cdot}9$ of the r.m.s. value, but with other waveforms the factor is in general different, so if (as is usual) the instrument is scaled in r.m.s. values it must be remembered that there may be quite large errors if the waveform is not sinusoidal. In *J. Sci. Inst.*, Sept. 1948, A. Cunliffe shows that in the worst phase conditions the percentage error caused by an odd harmonic is approximately equal to the percentage harmonic divided by the number of the harmonic, but the error due to an even harmonic is less serious, being approximately equal to the square of the percentage harmonic divided by 200. Thus the worst error due to 15% second harmonic would be about 1·1%, but the same percentage third harmonic would cause anything up to 5% error. It is possible to combine rectifiers with resistors to make a nearly square-law voltmeter, in which waveform error is practically eliminated; see Sec. 5.14.

The top frequency of the early types of instrument rectifiers was about 5 kc/s, but this has been considerably raised and now includes at least the full a.f. range. Germanium rectifiers are being used for meters up to 100 Mc/s, and silicon rectifiers up to 10,000 Mc/s. Rectifier meters are considered further in Sec. 5.12.

The electrostatic instrument is a true voltmeter and takes no current, except a capacitive current that limits its usefulness at high frequencies. The force due to low voltages is so small that an electrostatic meter to read them is delicate and costly; but at high voltages, such as those used in television, for which current-operated voltmeters are least suitable, the electrostatic type comes into its own.

For further reading the following books are recommended: *Electrical Instruments and Measurements*, by W. Alexander (Cleaver-Hume); *Electrical Measuring Instruments*, by C. V. Drysdale and A. C. Jolley (Chapman & Hall).

5.3. ACCURACY OF METERS

Some meters are marked " B.S.1 " or " British Standard First Grade ". This does not mean that they are the highest possible grade of instrument, for there are better ones called " Sub-standard " (which does not sound nearly so impressive!). What it does mean is that the instrument satisfies a standard of accuracy, good enough for most ordinary work, laid down in the early editions of B.S. 89.* Because these descriptions were rather misleading they were altered in the 1954 edition to " Industrial " (" In ") and " Precision " (" Pr ") respectively. This publication specifies the construction and testing of the various types of meters in great detail, and the limits of error allowed in the two grades, over the effective range, measured at standard temperature (20° C). Allowable temperature errors are additional. The effective ranges are reckoned to be from full-scale value (i.e., the higher of the values associated with the ends of the scale) down to $\frac{1}{10}$ of full-scale for moving-coil, $\frac{1}{5}$ for moving-iron, $\frac{1}{4}$ for rectifier, and $\frac{1}{4}$ for electrostatic and thermal types, except that in electrostatic voltmeters having smaller than $3\frac{1}{4}$-in scales it is from $\frac{1}{4}$ to $\frac{3}{4}$ scale-range (i.e., the number of units between the two scale-end marks).

The limits of error, so defined, for Precision single- or multi-range voltmeters or ammeters (except self-contained multi-range ammeters) is 0·3% of the scale range for moving-coil and 0·5% for all other types. For self-contained multi-range ammeters it is 0·5% for all types. Rectifier types are not regarded as accurate enough to come within the Precision grade.

The maximum errors for Industrial instruments are specified in Table 5.1.

There is an extra allowance of 0·25% for multi-range instruments, and 1% for moving-iron meters used on d.c. The allowable temperature error per °C above or below 20° C (35° C in tropical models) is tabulated in B.S. 89: 1954 and varies from 0·03 to 0·1% for Precision meters and 0·10 to 0·30% for Industrial.

E.g: If a portable multi-range moving-coil volt-milliammeter with a 4-in scale were reading 15 V on its 100-V range, the maximum error at 20° C within the Industrial rating would be $1{\cdot}5 + 0{\cdot}25\%$ of 100 V, so the true voltage might be anything from 13·25 to 16·75. If used in a freezing-cold situation (0° C) the tolerance would be widened by another $0{\cdot}15 \times 20 = 3$ volts. So even an up-to-standard moving-coil instrument with a reasonably large scale can be considerably in error if used in unsuitable conditions. Although the effective range is reckoned down to $\frac{1}{10}$ full scale, it is advisable to use the upper half or at most two-thirds of the scale if at all possible, and not to try to make do with a barely sufficient number of ranges. Temperature errors can be serious; even more so with rectifier and thermal meters. And in low-power circuits the power drawn by even the best moving-coil meters ought not to be neglected. The voltage drop on the current ranges (standard 0·075, but 0·1 or even higher is common) may occasionally have to be allowed for, especially in transistor circuits; and the current taken during voltage readings (Sec. 10.1), especially in

* *Electrical Indicating Instruments* (B.S.I.).

valve circuits. The latter correction is easier if one of the scales shows the current taken by the instrument as a voltmeter.

Another kind of error can arise, especially in moments of stress, if the factors by which the scale readings have to be multiplied are awkward ones like 4 and 2½. The only ones permitted by B.S. 89 : 1954

Table 5.1

LIMITS OF ERROR FOR INDUSTRIAL GRADE INSTRUMENTS OVER THEIR EFFECTIVE RANGE, EXPRESSED AS A PERCENTAGE OF SCALE RANGE, AT 20° C

Type of Movement	*Current (I) or Voltage (V)*	*Portable (P) or Switch-board (S)*	*Scale length in inches*			
			5 *or over*	3½ *up to but not including* 5	2 *up to but not including* 3½	1¼ *up to but not including* 2
Permanent-magnet moving-coil	I or V	P	0·75	1	1·5	2
	I and V	P	0·75	1·5	2	—
	I or V	S	1	1	2	2·5
Moving-iron	I or V	P	1	1	2	—
	I	S	1·5	1·5	2	2·5
	V	S	1	1	2	2·5
Rectifier or Thermo-couple	I or V	P or S	2	2	2·5	3
Electro-static	V	P or S	1·5	2	2·5	—

are 1, 2 and ½, combined if need be with multiples or submultiples of 10. The present tendency is to provide two scales, which by the use of decimal multipliers give ranges in the series 1, 3, 10, 30, etc.

For higher accuracy than is possible with pointer instruments, see Sec. 5.10.

5.4. MULTI-RANGE METERS

For general purposes, multi-range meters are indispensable, and there is a wide choice. Almost all read direct current and voltage; most of them have alternating-voltage ranges; many read alternating current; and often they have some resistance, capacitance, etc., ranges. The nucleus is a moving-coil meter with a number of internal shunts to extend the basic current range (commonly 1 mA), and series resistances (multipliers) to provide voltage ranges. If still higher ranges are needed, they can be obtained by external shunts and multipliers. Alternating ranges necessitate a rectifier. For the alternating-voltage ranges the same multipliers can serve, if the sensitivity of the meter is increased 10% on those ranges for the reason explained in Sec. 5.2. Because of the non-linearity of the rectifier, low-reading a.v. ranges need a special scale; and for the same reason a.c. ranges cannot be provided by shunts but need a current transformer. Resistance ranges are provided as described in Sec. 5.8.

A well-known example of this class of instrument is the Avometer Model 7, having 50 ranges self-contained: up to 10 A and 1,000 V (d.c. and a.c.), 40 MΩ, and 0·01 μF. The " Multiminor " by the same maker has fewer ranges but is more compact, takes only one tenth of the current (0·1 mA full-scale), and achieves reasonable accuracy and exceptional reliability at low cost by the technique of printed resistors, as shown in Fig. 5.1.

In electronic circuits, even 0·1 mA is sometimes more than can well be spared for a voltmeter, and there are instruments (including an Avometer, and the Salford " Selectest Super 50 " and " Minitest ") taking only 50 μA. The Taylor Model 100 A goes even further, with a full-scale current of 10 μA, and measures up to 200 MΩ without an external battery. This high sensitivity is due in part to the permanent magnet being inside the moving coil. Two features often found in high-sensitivity meters are: (1) a mechanical cut-out to disconnect the

Fig. 5.1—Interior of " Multiminor " multi-range meter, showing printed range resistors. (AVO Ltd.)

instrument automatically if an excessive current is passed through the moving coil, and (2) a reversing switch to avoid having to transpose the external connections—perhaps in contact with a dangerous voltage.

The foregoing instruments come, at best, into the Industrial BS grade; the Precision grade naturally costs more. Most Precision meters are single-ranged, but an example of a multi-range instrument in this grade (on its d.c. ranges) is the Electronic Instruments Model 44. It has 43 ranges, reads up to 10 A and 1,000 V (d.c. and a.c.), has a 1 mA movement and 0·2 V drop on direct current ranges. As the BS specifications do not include a Precision grade for rectifier-type meters, the a.c. ranges are nominally Industrial, but with a good margin of accuracy.

In theory, the construction of shunts and multipliers for providing specified ranges is simple; in practice it demands more care for a given standard of accuracy than the uninitiated might suppose. The making of a universal shunt is described by A. L. Chisholm (*W. W.*, Jan. 1954). The whole subject of a.c./d.c. test meters, with instructions for making them, is covered by a book of that name by W. H. Cazaly and T. Roddam (Pitman). And as for buying a multi-range meter, one would be well advised to pay less attention to the number of ranges than to the quality and reliability of the construction.

5.5. RECTIFIER SHUNTS

Milliammeters—and, still more, microammeters—are subject to the occupational hazard of bent pointers, or a worse fate, as a result of accidental excess currents. The more sensitive multi-range meters, as just mentioned, are fitted with a re-settable mechanical cut-out. For instruments that are not, protection can be provided by a rectifier, which functions as an automatic shunt, its resistance falling steeply as the current increases.

Two modes of use must be distinguished, according to whether or not the calibration of the instrument is affected by connecting the shunt. Protection can be combined with a very wide-range approximately logarithmic scale (say, from 0·1 to 250 mA), and this type especially

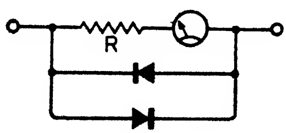


Fig. 5.2—Essentials of rectifier shunt system for protecting meters against excessive current and/or providing them with non-linear wide-range scale

was discussed by the author (*W. W.*, 11th Jan. 1935 and 2nd June 1938). There are two disadvantages: the calibration depends very largely on the rectifier, which varies with temperature; and, in order to divert a large proportion of the current through the rectifier at full-scale reading, the resistance of the meter must be increased to bring the voltage drop up to about 0·35 V or more. Copper oxide has the lowest " forward drop "; germanium is next best; selenium and silicon need about twice the voltage. However, for obtaining approximate readings of currents that are liable to vary widely, the idea is useful. Fig. 5.2 shows the circuit, in which *R* is the added resistance. The second rectifier diode is needed to protect the meter against reverse connection. Fig. 5.3 shows a typical scale compared with the original 0 to 5 mA calibration, using a copper-oxide rectifier with the lowest voltage rating, and *R* about 70 Ω.

The idea is even more applicable to instruments such as null indicators, which are not needed to give actual readings but must be sensitive near zero and not damaged by very much stronger currents.

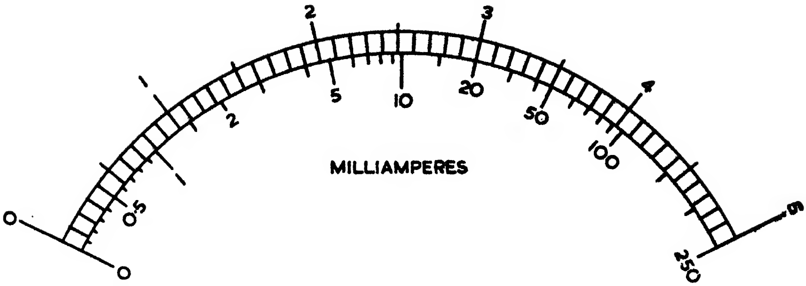


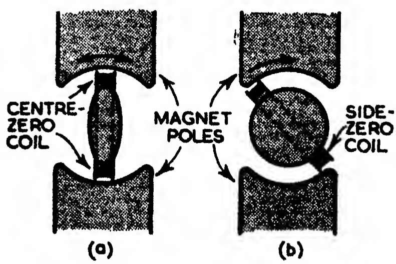
Fig. 5.3—Scale of rectifier-shunted milliammeter compared with original to show the great increase in range of current measurable, and the logarithmic scale shape

The commonly-used germanium diodes are suitable, choice of type depending on the maximum current that can flow. An alternative method of protecting the pointer of a moving-coil galvanometer is to use a core shaped something like Fig. 5.4*a*, giving a magnetic field that falls off steeply as the coil is deflected, instead of the usual shape (*b*) designed for uniform strength at all coil positions and hence a linear scale. This scheme does not, of course, protect the coil from being burnt out.

If the calibration is not to be affected by the shunt, the resistance of the meter is preferably low, and diversion through the shunt is much less. However, as J. de Gruchy shows in an informative article on rectifier shunts (*W.W.*, Sept. 1953) some instruments can stand overloads of the order of 10,000%.

The risk to a sensitive instrument is especially acute in high-voltage tests, but as there is then no need for its resistance to be low it is

Fig. 5.4—Shape of flux-concentrating iron core for (a) *null-indicating galvanometer, which is required to be sensitive only near zero, compared with that in* (b) *conventional linear-scale instrument*



possible to give very good protection by means of a single Zener diode, as in Fig. 5.5. By suitable choice of R_1 the diode can be arranged to have no effect on the calibration and yet break down completely at, say, only 50% above full-scale current. The meter is completely protected against reverse currents, for the diode is then in its "forward" direction. R_2, which may be wholly or partly the resistance of the source, protects the diode. It does not protect either the voltage source or whatever is connected to the test terminals.

What is done about that depends on the design and maximum allowable current of the source. There are, for example, fuses blowing at as little as 10 mA, and there is the device shown in Fig. 4.9.

Yet another device that can be used for instrument protection is the transistor. Some circuits have been given by G. G. Yates (*E.E.*, March 1960, p. 172).

5.6. WATTMETERS

Electrical power is the product of current and voltage, so one method of determining it is to measure or calculate each separately and multiply. That is all right for d.c., and for a.c. in non-reactive circuits, but if current and voltage are not in phase an extra factor must be included—the cosine of the phase angle, otherwise known as

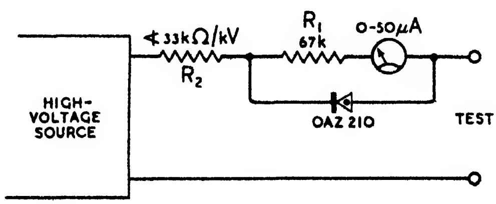


Fig. 5.5—Zener diode shunt for protecting microammeter used to test current leakage at high voltages

the power factor. A wattmeter, which depends for its deflection on the force between a movable and a fixed coil, one for voltage and one for current, takes account of this and reads true watts. Although wattmeters are very important in " heavy " electrical engineering, they are not much used in electronic laboratories, because they are comparatively expensive, are subject to considerable error due to their own consumption in low-power systems, and are particularly inaccurate at the low power factors common in communication circuits. Except perhaps for measuring the total power taken by a receiver there are other ways (Secs. 10.6 to 10.7) of measuring power factor.

The foregoing are the fundamental pointer instruments; there are others, going under various names, in which the same meters appear thinly disguised by scales calibrated in other quantities. As wattmeters are still fresh in the mind this is perhaps a good point at which to expose the pseudo-wattmeter, which is simply a voltmeter or milliammeter calibrated in watts. Such a calibration can hold good only under certain conditions: for instance, an ammeter may be connected in a supply circuit, and assuming a certain fixed voltage and power factor it is legitimate to scale it in watts. But of course its readings are immediately invalidated if voltage or power factor alter; unlike a true wattmeter it cannot take account of these variations.

If the circuit in which the power is to be measured consists of a pure resistance, the power factor need not be taken into consideration because it is equal to 1. The power is equal to E^2/R or I^2R, so a voltmeter or ammeter can be legitimately scaled to read watts consumed in a particular resistance. This is the principle of the output meter, which, being rather important in radio work, deserves more than passing mention.

5.7. OUTPUT POWER METERS

When testing a receiver by applying a signal, the ear is an unsatisfactory indicator because it cannot easily distinguish small changes in intensity of sound, it gives no quantitative readings, and it cannot be relied on even for rough comparison. So for comparative work—e.g., " lining up " a receiver—an ordinary metal-rectifier a.c. voltmeter can be connected across primary or secondary of the output transformer (preferably the latter if the meter possesses a suitable range). A valve voltmeter is sometimes recommended for this purpose, but is much less convenient than the metal-rectifier type.

Now a loudspeaker is not a fixed pure resistance, but, if its resistance as a load at the frequency of test is known at least approximately, some idea of the actual power in it can be inferred from a voltage reading. Of course it is necessary to make sure that the resistance of the voltmeter is at least several times as great as that of the speaker!

When more accurate measurements of power are wanted, as in determining the sensitivity or maximum power output of a set, it is better to replace the loudspeaker by an artificial load consisting of an accurately known pure resistance. The voltmeter can then be scaled in watts or milliwatts.

While the instrument is certainly capable of reading true watts in this way, it is taken for granted in measurements of the output of transistors or valves that the load resistance is adjusted at least approximately to the optimum value, or whatever other value may be required for special tests. Not only is it necessary, then, to have a load resistance that can be varied to cover all probable requirements, but the milliwatt scale of the voltmeter has to be different for every value of resistance, which is hardly practicable if a really adequate selection of resistances is provided.

One way of avoiding this difficulty is by means of a transformer with a secondary winding across which the voltmeter and a suitable load resistance are connected. The primary has a considerable number of tappings so that the fixed load resistance can be made equivalent to as many different resistances looked at from the primary side. To present a resistive impedance over a wide range of values at all audio frequencies is a difficult transformer design problem. The Marconi Instruments TF893 provides as many as 48 impedances from 2·5 Ω to 20 kΩ by combining primary tappings (a decade multiplier) with a switched resistance network on the secondary side. There are five power ranges, from 1 mW to 10 W full-scale and the effective frequency range is 50 c/s to 10 kc/s, or 20 c/s to 35 kc/s with reduced accuracy.

The wider frequency range of 10 c/s to 250 kc/s is obtained in the Heathkit Model AW-IU by using a purely resistive switched network in place of the transformer, but the choice of load resistance is restricted to four values: 3, 8, 15 and 600 Ω. Owing to the loss in the network, a two-stage amplifier with negative feedback is needed to operate the voltmeter. A variable resistance meter is very valuable, for it enables one to see quickly how the output of a stage is affected by the load resistance. The resistance which is found to give maximum output is not, of course, generally the optimum load; one has to take account of distortion too.

5.8. OHMMETERS

Another popular meter scaled to read an indirect quantity is the ohmmeter. Most multi-range test sets are provided with a direct-reading resistance scale. The power for deflecting the pointer is obtained from a battery—usually a small one contained in the instrument, to be seen for example in Fig. 5.1—and the extent of the deflection is controlled by the resistance connected to the appropriate terminals or clips. As the deflection depends also on the voltage of the battery a preliminary adjustment is necessary to compensate for this.

Fig. 5.6 shows the circuit of one of these simple ohmmeters. R is a resistance equal to the desired mid-scale reading, and the meter is

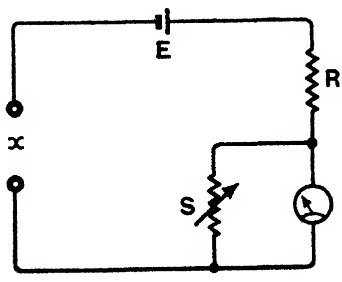


Fig. 5.6—Circuit of simple ohmmeter. The resistance to be measured is connected to the "x" terminals

adjusted by the variable shunt S to give its full deflection when the x terminals are short-circuited. The maximum variation in resistance of meter and battery is assumed to be negligible compared with R, and their normal resistance is included in R for purposes of calculation. Under these conditions I_1, full scale current, $= E/R$. The resistance,

R_x, to be measured is then connected and the current now flowing, I_2, $= E/(R + R_x)$. Eliminating E we get

$$R_x = R\left(\frac{I_1}{I_2} - 1\right),$$

from which it is seen that the result depends on R (which is fixed) and the *ratio* of the two currents. As the actual values of current therefore need not be known it is allowable to use the variable shunt, as described, for bringing the pointer initially to full deflection (zero on

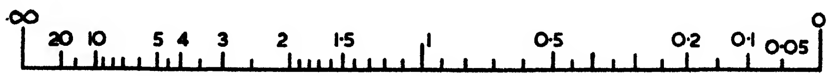


Fig. 5.7—Scale of the Fig. 5.6 type of ohmmeter, requiring only to be multiplied by R to indicate the resistance connected at " x "

the resistance scale) whatever the battery voltage may be; the second deflection, with R_x in circuit, is then a measure of its resistance.

The Avometer, a pioneer in this field, has *two* preliminary adjustments—one compensating for changes in voltage and the other for battery resistance.

Fig. 5.7 shows I_1/I_2—1 plotted along a linear 0 to I_1 scale, and is a universal resistance scale for this type of instrument, the reading only needing to be multiplied by R to give R_x. Theoretically, all values of resistance are included on such a scale, no matter what the value of R; but those that can be clearly read are limited to the range $R/10$ to $10R$. By selecting suitable values for R and the shunt, several overlapping ranges can be provided in one instrument. For information on how to make multi-range ohmmeters, see F. L. Hogg (*W.W.*, Aug. and Sept. 1943); Cazaly and Roddam's *A.C./D.C. Test Meters* (Pitman); and W. Tusting (*W.W.*, July 1953).

Even with more than one range it is difficult with one instrument to cover all the resistances encountered in electronic practice. The conventional ohmmeter seldom has enough voltage to go up into the megohm ranges without additional batteries, or enough current for the very low resistance ranges. It is often desirable to have some means of measuring the latter—switch contacts, etc. A simple method is to use the unknown resistance as a shunt to a current meter, the reduction in deflection being a measure of the resistance. To cover a wide range of low resistances in one instrument the author devised a multi-ranging extension of this scheme, which has been embodied in the Cambridge " Resistance Meter ", shown in Fig. 5.8, as one of the many accessories available for use with the Cambridge " Unipivot Versatile Galvanometer " to which it is attached. The same scale applies for all four ranges, covering from 0·001 to 1,000 Ω with reasonable accuracy and the scale factors are all multiples of 10. There is no possibility of bumping

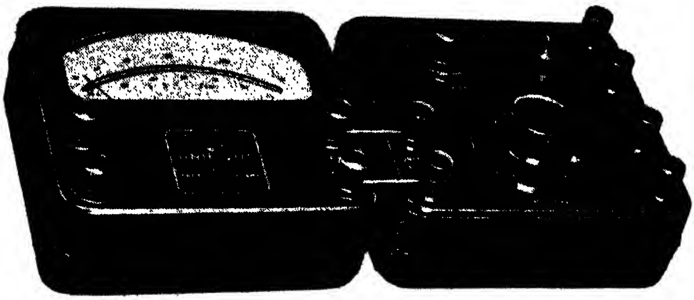


Fig. 5.8—The " Unipivot Versatile Galvanometer " can be used for accurate work in conjunction with a great variety of accessories. A multi-range ohmmeter manufactured to the author's design is shown. (Cambridge Instrument Co., Ltd.)

the pointer, even if the unknown resistance fluctuates between zero and infinity, as may approximately happen when one is joggling a switch to test its contacts. The voltage employed (2 max.) is sufficiently low not to give a misleading indication by breaking through surface skins.

But when testing the *absence* of a conducting path it is usually desirable to apply a *high* voltage, to ensure that the insulation under test is capable of withstanding it. For this the Megger, made by Evershed & Vignoles, is the accepted instrument. As high-resistance testing at high voltage forms a relatively small part of the work of most radio laboratories, the cost of the full-sized Meggers puts them out of reach of all except rather lavish establishments, but the " Megger Tester Series 3 " illustrated in Fig. 5.9 and 10 is relatively inexpensive. A typical machine generates 500 V and is scaled from 10 kΩ to 50 MΩ. The reading is independent of the speed at which the generator handle is turned, provided that the capacitance of the circuit tested does not exceed about 4 μF. Other models are obtainable with a choice of scales up to 200,000 MΩ at 10,000 V, if need be. When it is not necessary to be independent of an electricity supply, the testing voltage can be generated in the usual way with a step-up transformer and rectifier.

For the measurement of still higher resistances by means of valves, see Sec. 5.18.

For installation testing it is very helpful to have a portable instrument with both megohm and low ohm ranges. The Everett Edgcumbe " Metrohm ", which is energized by an internal 9-V battery, has one range of 0·1 to 100Ω, using a 1·3 mA meter in a Fig. 5.6 type of circuit; throwing a switch connects the battery to a transistor voltage transformer yielding (in one model) 500 V in series with the meter to measure 0·15 to 50 MΩ. A printed circuit is used.

5.9. VIBRATION GALVANOMETERS

Although not exactly a pointer instrument, the vibration galvanometer may be mentioned here. It can be either of the moving-coil

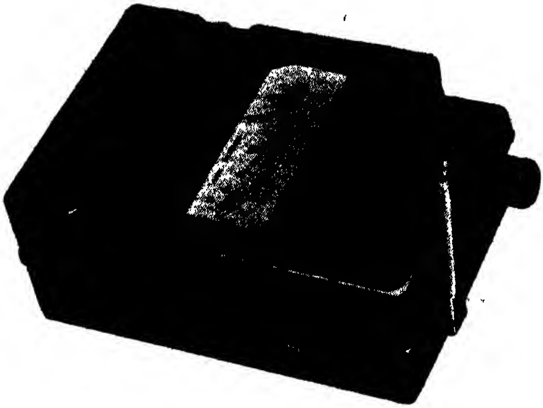


Fig. 5.9—Insulation and Continuity Tester Series 3, a direct-reading ohmmeter incorporating a hand-driven generator. High and low resistance scales are selected by the switch on the side

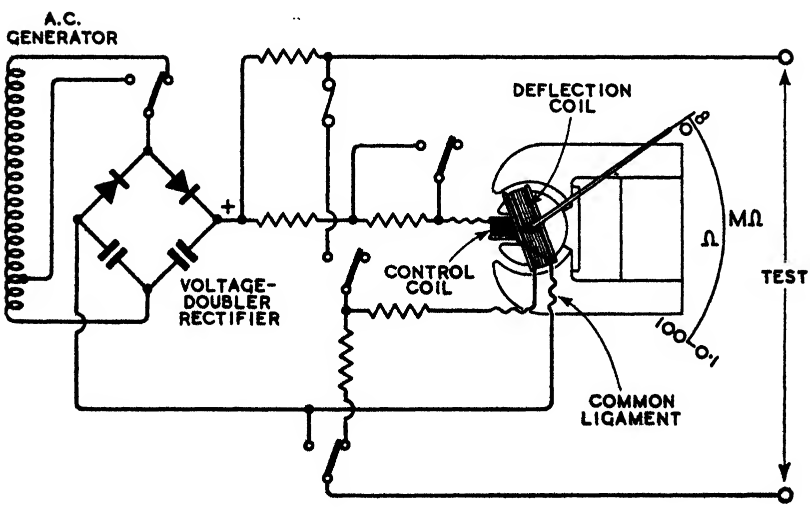


Fig. 5.10—Circuit of the Series 3 Tester. With switch arms in the positions shown it is connected for insulation testing. With infinite resistance, current flows only through the control coil, causing the pointer to be held in the position shown. Any current flowing through the resistance under test traverses the deflection coil and causes a corresponding deflection. For continuity or low-resistance tests, the switch arms are to the left and the deflection coil is in parallel; the applied voltage is reduced from 500 to about 4

or moving-iron type; if the former, the word " coil " must be interpreted rather broadly, as the moving element usually consists of a fine wire through which the signal current passes between the poles of a permanent magnet. If the current is alternating, the moving element vibrates, and as it has a very sharp mechanical resonance the response is highly sensitive to signals of a particular frequency. The vibration is made visible by a lamp and lens system. The principal application is to bridge work at frequencies of the order of 50 c/s, at which it is extremely sensitive and has the great advantage of practically ignoring harmonics and other currents that might tend to obscure the null point. Vibration galvanometers are usually fairly expensive; an alternative with a rather wider range of usefulness is a sharply tuned amplifier with meter or other indicator (Sec. 5.23). Both the frequency and sharpness of tuning can readily be adjustable over wide ranges.

5.10. DIGITAL METERS

The accuracy of a pointer type of meter is strictly limited. As can be gathered from Sec. 5.3, it is only under favourable conditions that measurements with the everyday quality of instrument can be relied upon within 2%. The cost of higher precision increases more rapidly than inverse proportion to the error, and tends to infinity in the region of 0·1%. Some care and skill is needed to make precise readings. Greater accuracy and reliability are obtained by potentiometer measurements (Sec. 10.2) but they require even more skill and time. By the use of digital computor techniques, however, measurements can be made automatically with great accuracy and speed, and no skill at all. Another advantage is that the readings can be fed directly into a computor. An all-transistor digital voltmeter described by R. C. Bowes and J. C. Gill (*Proc. I.E.E.*, Vol. 106, Part B, Supp. 18, May 1959), being intended primarily for this purpose, works in binary digits. The methods adopted differ widely, but one thing they have in common is complexity and high cost. In general, the accuracy depends on that of the standards contained (usually voltage reference devices and wire-wound resistors) and on the way in which the quantity to be measured is compared with them.

A digital voltmeter described by H. Sutcliffe (*E. & R. E.*, May 1959) is in effect an electronically operated potentiometer, in which the voltage to be measured is compared with one developed in the instrument. The difference between them varies the internal voltage until it equals the external, and the answer is displayed to four figures on " Dekatron " tubes. The range of measurement is 0·1 mV to 10 V, the maximum error 0·02% of full scale (1 V or 10 V), the maximum time to make a measurement 1 second, and the maximum input current 0·001 μA.

Several types can be bought; in most of them, such as the Solartron LM901 and LM902, the voltage is displayed in figures. The latter measures up to 1,599 V in five ranges, to an accuracy of 0·1%, and is operated almost entirely by transistors. The same firm produces digital ohmmeters and bridges.

The principles and applications of digital measurements have been expounded by P. R. Darrington (*W.W.*, June 1961).

5.11. CLIP-AROUND METERS

Another device which provides great convenience where the nature of the work justifies the expense is the clip-around or clip-on meter, enabling current to be measured without breaking the circuit. A.c. presents no great difficulty, as the conductor carrying the current to be measured acts as the one-turn primary of a transformer, and the clip-on probe the core, on which a secondary is wound. Since the voltage generated is obviously very low, an amplifier is needed between the probe and meter, as considered in Sec. 5.22. Use of the probe increases the error somewhat—e.g., 5%, compared with 2% for the main instrument—and is liable to interference from stray a.c. fields, so provision for alternative connection across a low resistance in the circuit is usual. The Dawe Type 618 " Milliclamp " comprises a probe and an adapter for enabling it to be used with a valve voltmeter or oscilloscope. The current range is from a few hundred μA to 10 A, at frequencies between 7 c/s and 1 Mc/s. The Hewlett-Packard Model 456A probe has a frequency range of 25 c/s to 20 Mc/s.

For measuring d.c., its steady field around the conductor is applied in the probe to what is virtually a small magnetic amplifier. An a.c. signal fed to this is normally balanced out, but the magnetic field unbalances it and allows a proportionate alternating voltage to appear at the input to the amplifier driving the meter. Such an instrument is particularly useful for transistor work, where current rather than voltage is important and the insertion of a milliammeter is liable to change the conditions; and for computors, in which very many readings are required. Another useful feature is that the sum or difference of currents can be read by clipping the probe round two wires at once. This is a help in balancing push-pull amplifiers. It is a pity that the clip-on facility multiplies the price so many-fold.

The Hewlett-Packard Model 428A has six ranges, 3 mA to 1 A f.s.d., and an accuracy with 3% ± 0·1 mA. A.c. in the test circuit with a peak value less than full-scale affects accuracy less than 2%. The probe induces less than 15 mV a.v. and less than 0·5 μH into the test circuit. Fig. 5.11 shows the appearance. The Solartron AM1002 is a valveless instrument with similar ranges, and rather less reaction on the circuit—less than 0·5 μV and 0·05 μH.

5.12. METER RECTIFIERS

In the instruments just described, the indicators are assisted by amplification, which is much used for increasing sensitivity without making the meters more delicate. Those that are required to read a.c. or a.v. almost invariably use a moving-coil meter with a rectifier, as in Sec. 5.2. Before we go on to consider meters using amplification, the principles of rectifiers used for measurement—with or without amplification—deserve some more attention.

As we saw in Sec. 5.2, the matter is complicated by the question of waveform and "values". For many purposes one would like to know r.m.s. values, because power is proportional to the square of r.m.s. current or voltage, regardless of waveform. Unfortunately

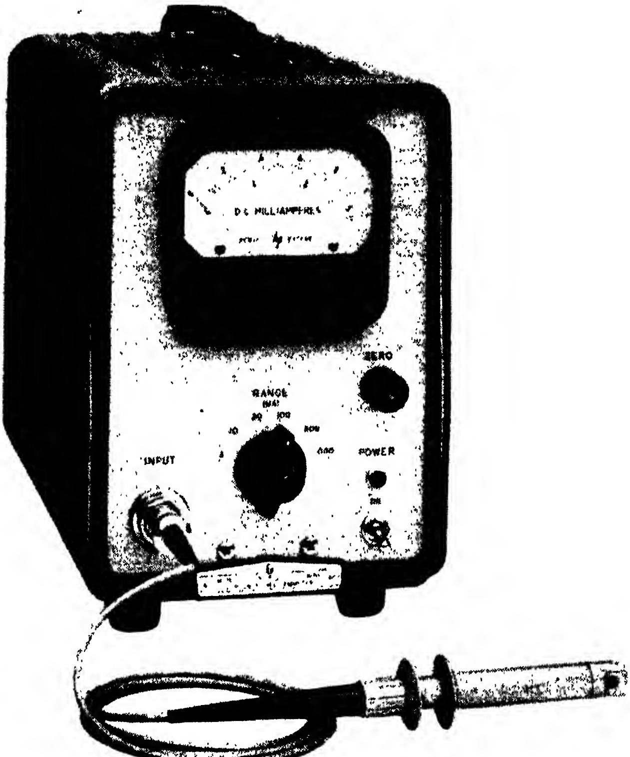


Fig. 5.11—Hewlett-Packard clip-around milliammeter Type 428A, by which d.c. can be measured without breaking the circuit. (Livingston Laboratories, Ltd.)

r.m.s. values are not shown by any simple and reliable rectifier type of meter. The tendency is to read either mean values (which are seldom of much interest) or peak values, according to arrangement. If the scale is then marked in r.m.s. values, assuming sine waveform, the readings with other waveforms may have substantial errors (Sec. 5.2). With the extensive use of non-sinusoidal waveforms, cathode-ray equipment has tended to displace valve voltmeters and other rectifier meters, so that waveforms can be seen as well as measured; but for many purposes a meter is more convenient.

A moving-coil meter in series with a perfect rectifier, no capacitor being in circuit, shows mean values. If the rectifier is half-wave, the means of positive and negative half-cycles can be read separately, by reversing connections; if full-wave, the sum of both is shown, whichever way the instrument is connected. For power and audio frequencies, the usual multi-range meter has a bridge-connected full-wave copper-oxide rectifier, as in Fig. 5.12, where it is arranged as a voltmeter, with multiplier R. Resistance should *not* be connected in

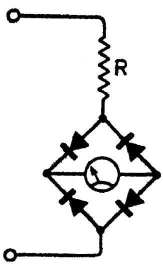


Fig. 5.12—Connection of moving-coil milliammeter for use as an a.c. voltmeter

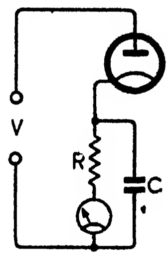


Fig. 5.13—Basic circuit of diode voltmeter responsive to nearly peak value

series with the meter itself, which should have a low resistance; the voltage drop across it should not exceed 0.5. The reason for this, and the bridge connection, is that the instrument type of rectifier cannot stand much reverse voltage.

Owing to the curvature of the rectifier characteristic, the linear part of the voltmeter scales is displaced from zero about 0·5 V. For the same reason, current ranges cannot be obtained in the usual d.c. manner by shunts; a current transformer is needed. Further information is given in Westinghouse Brake & Signal Co's Publication MR3.

Much higher frequencies can be covered by using suitable germanium or silicon diodes instead of copper-oxide, but the curvature extends to rather higher voltages.

The mean value of sinusoidal half-cycles is $1/\pi$ ($= 0{\cdot}318$) times the peak value, and this is what is indicated by a d.c. voltmeter in series with a single diode. By simply connecting a sufficiently large capacitance across the voltmeter, the deflection is increased three-fold to nearly the peak value. So unless there is any particular reason for preferring readings proportional to the mean value, this is commonly done, especially in valve voltmeters. Because the basic circuit (Fig. 5.13) looks so simple it tends to receive little consideration. But it is less simple than it looks, and after several decades of use is still often misunderstood. The following is an outline of the author's attempt

(*W.W.*, March 1952, and June and July 1954; mathematically, in *W.E.*, Feb. 1955) to clarify its working.

5.13. THE DIODE VOLTMETER

If the diode were a perfect rectifier and R were infinite (open-circuit), C would charge up to the peak value of V (V_{max}) and if there were no leakage would remain so indefinitely. Some leakage path (R) is necessary to enable the rectified voltage (V_d) to follow changes in V. V_d is then bound to fall somewhat between successive positive peaks, as in Fig. 5.14, and the shortage has to be made up by the source recharging C at these peaks. If the time constant CR is long compared with one cycle, and the resistance of the charging circuit (source of V, plus forward resistance of diode) is much less than R, the recharging is confined to a small fraction of the cycle near the peak, as shown. For example, if V_{max} is 10 V and R 100 kΩ the discharging current is nearly 0·1 mA; and if the recharging occupied one-twentieth of the cycle the charging current I would have to average $20 \times 0{\cdot}1 = 2$ mA over that time. So the resistance of the rectifier would be infinite for 95% of each cycle and only $10/2 = 5$ kΩ during the remaining 5%. The single constant resistance (R') that would run away with the same amount of power as the whole rectifier system can be calculated as follows. On the assumption just made, practically all the power dissipation is in R, and as the voltage across R is nearly V_{max} all the time this dissipation is V^2_{max}/R watts, $= 2\ V^2_{rms}/R'$. By definition of R', this is equal to V^2_{rms}/R', so $R' = R/2$.

Fig. 5.13 necessitates a d.c. path through the source, and as this is often lacking in a.c. circuits the rectifier is usually rearranged as in Fig. 5.15. What is often overlooked, however, is that R is now continuously across V, so R' is equal to $R/2$ in parallel with R; that is to say, $R/3$.

When the source of V is a sharply resonant circuit, its flywheel effect keeps the input waveform reasonably sinusoidal (Sec. 4.16) so notwithstanding the extreme unsteadiness of the rectifier load its general effect is the same as that produced by a constant resistance R'. But when the source is a non-resonant generator, such as the output of a valve with resistance coupling, the effect is quite different, because it has no stored power with which to meet the brief but heavy load during the recharging phase. Consequently the positive peak is much more heavily cut, which means not only waveform distortion but a corresponding drop in the indicated voltage. Fig. 5.16 shows how small the source-plus-diode forward resistance (denoted by r) has to be, compared with R, in order to avoid an appreciable drop in indicated voltage V_d compared with peak " signal " voltage, V_{max}. Even when r is only one-hundredth of R it causes an error of nearly 10%, and if it were equal to R the voltmeter would cut the voltage down to one-fifth!

The effects of this on valve-voltmeter design are far-reaching. Obviously in Fig. 5.13 R cannot be made enormously large if there is

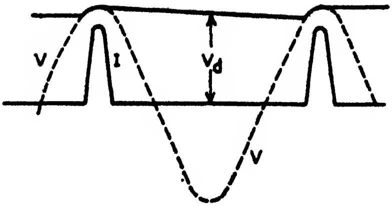


Fig. 5.14—Rectified voltage(V_d) and current (I) in relation to input voltage (V) in diode rectifier

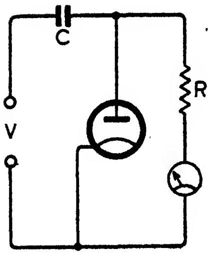


Fig. 5.15—Parallel form of diode voltmeter, which does not necessitate a conductive source

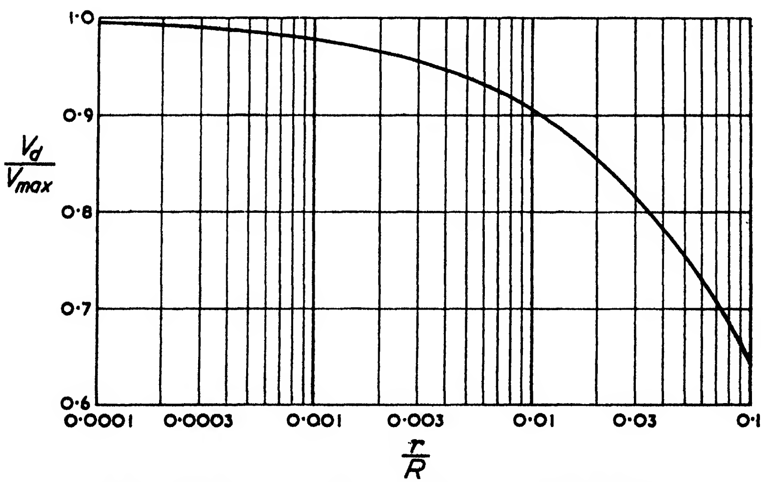


Fig. 5.16—Effect of the ratio r/R (input and diode resistance to load resistance) on efficiency of diode rectification

to be a reasonable meter current. Even when there is an amplifier between R and the meter, special precautions are needed to enable R to be as much as 1 MΩ without appreciable error due to amplifier input conductance.

Assuming 1 MΩ, however, Fig. 5.16 shows that an r of 3 MΩ causes 5% error, which on a 100-V range is much greater than that due to an ordinary 1,000 Ω/V metal-rectifier voltmeter! And the diode type needs an indicator 10 times as sensitive. So, for work on non-resonant circuits

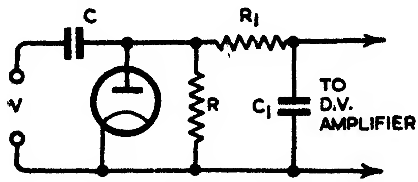


Fig. 5.17—The influence of the values of the components marked here on the performance of the diode rectifier is discussed in the text

at least, the common assumption that almost any valve voltmeter loads the circuit less than an ordinary voltmeter is ill-founded. And that is before even beginning to consider input capacitance.

In reckoning source resistance, the forward resistance of the rectifier itself cannot be left out of account. Although with an input of several volts it may be of the order of 500 Ω, at low voltages it rises to thousands of ohms, sufficient on its own to cause an appreciable discrepancy between output and peak input; in fact this is the cause of the well known non-linearity of the calibration towards the zero end—the " bottom bend " which cannot be got rid of, however large R is made.

It should be noted that the change from series (Fig. 5.13) to shunt (Fig. 5.15) causes V_d to be accompanied by the full alternating voltage V as well; and to prevent this from overloading any following amplifier and otherwise making a nuisance of itself it is usual to put a filter between diode and amplifier. What must not be overlooked is the effect of this on R'. In Fig. 5.17, if the amplifier can be neglected, the d.c. resistance remains as R, but the a.c. resistance is R in parallel with R_1; so R' is $R/2$ in parallel with R *and* R_1.

A higher value is obtained if R is connected across C_1, but V_d is reduced. (A $\sqrt{2}$: 1 reduction makes the output approximately equal to V_{rms}, assuming sinusoidal input.) Sometimes resistance is connected across C_1 without removing R; this arrangement reduces input resistance without any advantage.

Unlike other diode rectifiers, the thermionic type gives an output voltage with no input. In typical diodes it is roughly proportional to heater voltage (V_h), above half the rated voltage, and is fairly accurately proportional to the logarithm of R when R is not less than about 50 kΩ. A typical value of this zero displacement voltage for rated V_h and $R = 1$ MΩ is 0·5 V. There is little if anything to be gained by under-running the heater; the best remedy is to balance the undesired voltage out with that from another diode of the same type, which automatically preserves the balance when V_h varies.

One result of the bottom bend is that voltages much below 0·1 can hardly be measured at all. Another is that at least up to several volts one cannot use the same linear scale for all ranges. The only effective remedy is pre-diode amplification, which calls for skilful design to avoid drastic reduction in accuracy or frequency range or both. But it gets over the still more troublesome source-resistance difficulty. See Sec. 5.22.

A solution of the bottom-bend scale difficulty—but a very inconvenient one—is to use calibration curves. Another is to have separate scales for the non-linear ranges, but the poor meter may be already overloaded with scales for current, resistance, capacitance, etc. Perhaps the most convenient to use, though a little troublesome to provide, is a system which varies the filter resistances of the active and balancing diode so as to make all the linear parts of the curves pass through zero and have the correct slope. This leaves only one non-linear calibration, on the lowest a.v. range. The method is explained in the March 1952 *W.W.* article mentioned, and an example is given in Sec. 5.21.

At high frequencies the capacitance of the diode is important. For v.h.f., particularly, it is necessary to use a miniature uncapped diode such as the Mazda 6D1 (1·6 pF), mounted in the head of a flexible probe so that it can be brought right to the point where the voltage is to be measured. Even more important, perhaps, is to ensure that the input capacitance has as nearly as possible zero power factor, so that it does not substantially increase the power loss caused by the rectifier. C is usually a low-loss ceramic type, but at low frequencies its capacitance would be insufficient; for these the probe is arranged to plug into the main body of the voltmeter so that a larger C is connected in parallel (Fig. 5.26). To keep the error due to it below about 1 %, the time constant CR should be at least 20 times the period of the lowest frequency; say 1 (megohm-microfarads) for 20 c/s.

5.14. PEAK-TO-PEAK AND R.M.S. METERS

Unless one wants to measure the peak values of positive and negative half-cycles separately, a fairer single measure of (say) waveforms with even harmonics is given by a type of instrument that measures peak-to-peak values. This is done by the voltage-doubler rectifying circuit shown in Fig. 5.18, known also as the diode pump circuit. The effective input resistance R' is approximately $R/4$, but if R is doubled (as compared with Fig. 5.13) R' is the same and since the output voltage is also doubled the meter current is the same. In general, in order to avoid excessive source-resistance error and loading (discussed in Sec. 5.13) R would be too high for direct operation of a meter, and amplification is used, the voltage doubling then being directly advantageous. An example is the well-known " Heathkit " valve voltmeter, in which a double-diode valve works into a load resistance of 50 MΩ. C_1 and C_2 are 0·01 μF.

Thermionic diodes have several disadvantages: their relatively high forward resistance (r in Fig. 5.16, which tells the story), zero-input current to be balanced out, heater current, and size. Crystal diodes are more compact and convenient, have no zero-input current, and

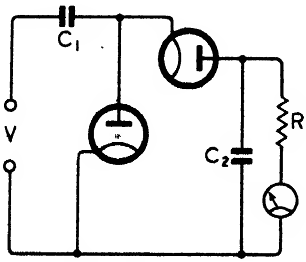


Fig. 5.18—Voltage-doubler or pump diode circuit, for measuring peak-to-peak values

can be obtained with very low forward resistance; but the advantage of low r tends to be offset by appreciable reverse current, which limits the possible R. Such diodes—junction types—also have a low upper frequency limit. But below 1 Mc/s or so they are usually to be preferred. Silicon is better than germanium for small reverse current and its relative indifference to temperature.

If there is a need to measure r.m.s. values regardless of waveform, the output to the moving-coil meter must be proportional to the *square* of the input. (The meter itself (aided by a capacitor) finds the *mean*.) And the square *root* is extracted by its non-linear scale.) The first of these three is the great problem. Certain valves, notably small triodes, have a square-law characteristic over a useful range, which can be found by plotting change in anode current against the square of the alternating grid voltage causing it, and noting to what extent the graph is a straight line. (The valve operates as an anode-bend detector. Both zero setting and calibration are greatly affected by variation in anode and grid voltages, anode-circuit resistance, and the valve itself) so this is not a convenient or reliable type of valve voltmeter, though it is capable of good results in skilled and patient hands. There are, however, more refined valve-characteristic methods; see R. M. Huey and J. E. Longfoot (*E.E.*, July 1960).

Another technique is to build up an approximation to a parabolic characteristic curve with a number of more or less straight sections, brought in successively by biased crystal diodes as the input voltage rises. Fewer of these are needed to give an acceptable approximation than one might expect; for instance, three diodes for a maximum error of about ±2%. Fig. 5.19 shows the type of circuit and characteristic used in the Bruel & Kjaer A.F. Spectrometer.* In general, the

* *B. & K. Technical Review*, July 1958.

relatively low input impedance—as well as the demand for millivolt ranges—necessitates prior amplification, which brings such instruments into the next Section; but reference may be made here to the Dawe Type 612 R.M.S. Valve Voltmeter which makes use of this kind of diode curve-shaping system. See also T. Roddam (*W.W.*, Aug. 1951) and the relevant part of a valve-voltmeter review (*E.T.*, Mar. 1960, p. 101).

Yet another technique is to use a meter of the thermal type (which is inherently r.m.s. responsive) preceded by a stable amplifier, as in the Cossor Model 1453, which has nine ranges, from 31·6 mV to 316 V f.s.d., at 50 c/s to 10 Mc/s. The Muirhead D-930-A precision r.m.s. decade voltmeter (*Muirhead Technique*, Jan. 1961) uses a lamp bridge as a null indicator, and a decade attenuator and range switch as the indicating elements; the frequency range is 5 c/s to 100 kc/s, and the

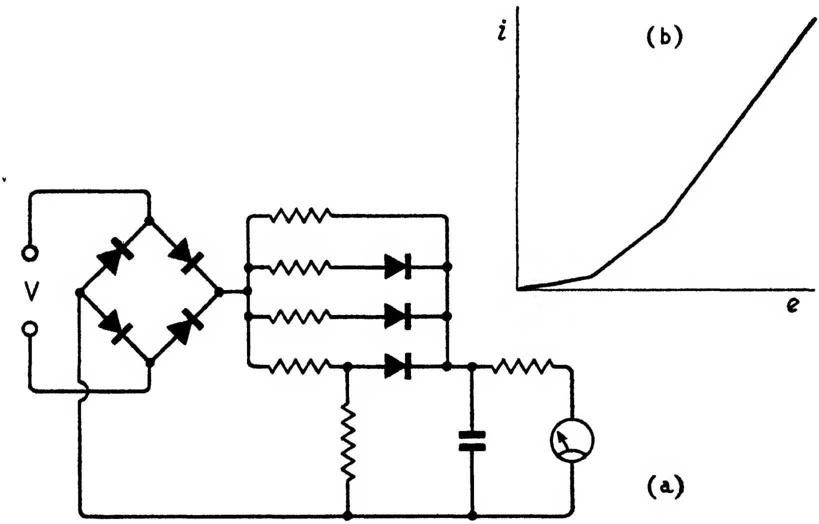


Fig. 5.19—(a) *Circuit and* (b) *approximation to square-law characteristic curve of rectifier meter using law-shaping diodes and resistors*

instrument is in a class by itself for accuracy, the error being of the order of 0·05%.

5.15. AMPLIFIER-AIDED METERS

The two main reasons for using amplifiers have just been mentioned: for increasing the sensitivity; and for reducing the effect of connecting the instrument, by substituting a high input impedance for that of a meter or rectifier. These advantages are at the cost of increased liability to error on account of possible variations in gain; reduced convenience; and, of course, money! The first of these is minimized by use of negative feedback; the second, by transistors, which help to free instruments from bondage to the mains, and reduce the bulk; the third seems to be left largely to the growth of prosperity.

For measuring direct voltage, obviously the meter is preceded by a stable d.c. amplifier. For alternating voltage there is a choice of arrangement: a d.c. amplifier following the rectifier; an a.c. amplifier preceding it; or both. (If voltmeters and voltages are mentioned most of the time, it is because current is usually measured by the drop in voltage across a known impedance.) The difficulty about d.c. amplifiers is that there is no very simple and effective way of distinguishing current changes due to a small voltage being measured from those due to other causes such as changes in valve characteristics, feeds, and temperature. Much progress has been made in the design of stable d.c. amplifiers, but designers still sometimes prefer to dodge the difficulties by converting the d.c. signal to a.c. and amplifying that. Such amplification is easy, because it is at a fixed frequency of one's choice.

Amplification of a.c. (i.e., before rectification) is especially attractive in a.c. voltmeters, because the inevitable non-linearity of even the best rectifiers causes increasing difficulty with scales as the voltage to be measured decreases, and makes reasonable accuracy difficult much below 1 V, and any measurement at all almost impossible at a few millivolts. Moreover, a rectifier is not a good kind of load to connect directly to a circuit (Secs. 5.13; 10.1). But however easy a.c. amplification may be at a fixed frequency or over a moderate range of frequency, it is difficult to keep constant from say 20 c/s to many Mc/s. This, then, is the chief problem in pre-detector amplification.

5.16. RESISTANCE TRANSFORMERS

An amplifier is of great value to a voltmeter even if it gives no voltage amplification at all. Ideally, a voltmeter would take no current from the circuit under test. But the most suitable type of indicator does take current. What is needed is an intervening unit, having infinite input resistance, zero—or known and constant—output resistance, and a constant voltage amplification ratio, even if it is only 1 : 1. It is the fact that valves can readily be arranged to approximate to this ideal that makes them so valuable for the purpose. It will be seen that such arrangements function primarily as resistance transformers rather than amplifiers. Though they necessarily give a large current amplification, their precise value is not a significant parameter, so the description " current amplifier " is hardly appropriate.

The disadvantages of a single valve in this role are that when working reasonably linearly there is a large initial anode current to flow through the meter, and that this current varies with valve and feed variations. The first trouble can be overcome by using the valve as one arm of a resistance bridge; and, as both can be largely overcome by using a similar valve as one of the other arms, a twin-valve arrangement, one variety of which is sketched in Fig. 5.20, is the basis of most valve voltmeters.

The relative merits of this (and the alternative with load resistors and meter on the anode side) and the corresponding single-valve voltmeters, are discussed in a valuable article by P. Popper and G. White (*W.E.*, Dec. 1948). See also J. D. Clare (*W.E.*, July 1948) on the Fig. 5.17 type. The second valve does nothing to increase the sensitivity,

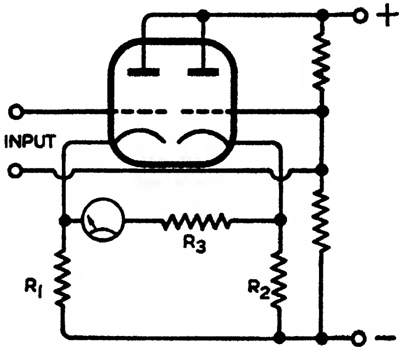


Fig. 5.20—The cathode-loaded form of the twin-triode d.v. amplifier commonly used in valve voltmeters

reckoned as meter-current change per volt input change; in fact, it halves it. But by automatically balancing out most of the undesired variations it enables a more sensitive meter to be used, thereby increasing sensitivity in terms of deflection. The advantage of cathode connection is the much lower output resistance than with anode connection. Working as a cathode follower, each valve has an output resistance of about $1/g_m$, which with a valve of, say, 5 mA/V is 200 Ω. R_1 and R_2 are assumed large in comparison. The resistance in series with the meter therefore consists of its range-adjusting multiplier, R_3, and nearly twice the output resistance of each valve. On the high ranges the latter is only a small proportion of the whole, so variations in it do not substantially affect the calibration; but if a low-reading range is used a major part of the meter-circuit resistance is " electronic " rather than the more reliable wire-wound sort. In the circuit shown, if R_1 and R_2 are large, the anode current—and hence g_m and the valve's output resistance—is determined almost entirely by R_2 and the right-hand grid voltage with respect to supply-negative, so it is sufficient to stabilize that grid voltage.

The need for any stabilization can be avoided if the amplifier has so little output resistance that variations in it have negligible effect on the calibration. Special types of triode such as the E88CC have g_m as high as 12·5 mA/V, corresponding to less than 160 Ω total series valve resistance in this circuit. But it is just the high-g_m types of valve that least fulfil the other ideal—infinite input resistance. It is not always realized that merely applying negative bias does not necessarily make grid current negligible. Comparison of anode current with and without a high resistance—say 50 MΩ—between grid and bias demonstrates this clearly. (The maker's *maximum* working grid-to-cathode circuit resistance for the E88CC valve is 1 MΩ !) As much as 1·5–2 V negative bias is needed to cut off ordinary grid current; beyond that there is current of opposite sign from several causes. For example, if the anode is at 100 V, anode-to-grid leakage through 100,000 MΩ is enough to make the grid 0·2 V more positive when 20 MΩ is in the grid lead! Ordinary valve holders are useless in such circumstances; materials such as PTFE are recommended.

There are special electrometer valves such as the ME1400 designed for extremely high input resistance. Without going to the slight extra expense of these, one can get results several orders better than usual by choosing a type such as EF86 which is known to have exceptionally low input conductance when run with low anode and heater voltages.* With suitable precautions the grid current of this valve can be kept below 0·001 μA.

These conditions are exactly opposite to those for low output resistance and a robust inexpensive meter. The answer is to use a two-stage amplifier, with appropriate input and output valves. The greater amplification presents the opportunity for greater negative feedback and consequent stability. To maintain the highest standard of performance a second similar pair is needed as a balance.

5.17. A HIGH-STABILITY VALVE VOLTMETER

A circuit devised by the author, very similar in appearance to certain others of the foregoing general description, but differing in important details, approximates very closely to the ideal by combining:

(1) Input conductance less than 10^{-10} mho (with under-run EF86 valves), which could be reduced still further if necessary by using electrometer valves;

(2) Output resistance less than 5 Ω, i.e., negligible even on a 0 to 1·5 V range using a 0 to 3 mA movement;

(3) Output voltage equal to input voltage within a fraction of 1 %, for any meter resistance from 300 Ω to infinity; so that the amplifier can be used as a resistance transformer for any voltmeter without the need for recalibration.

* "H.F. Pentodes in Electrometer Circuits", by K. D. E. Crawford. *E.E.*, July 1948.

"Receiving Valves Suitable for Electrometer Use", by G. A. Hay. *E.E.*, July 1951.

"Pre-amplifier Pentode, Type EF86", by W. A. Ferguson. *M.T.C.*, May 1954.

There is also the unconventional solution of using the anode as the input electrode and the grid as the output, with a transistor as current amplifier. No h.t. is required, and d.v. ranges from 0·3 to 600 V f.s.d. are feasible:

"Inverted Triode Voltmeter", by R. B. Rowson and A. P. Williams. *W.W.*, Aug. 1960.

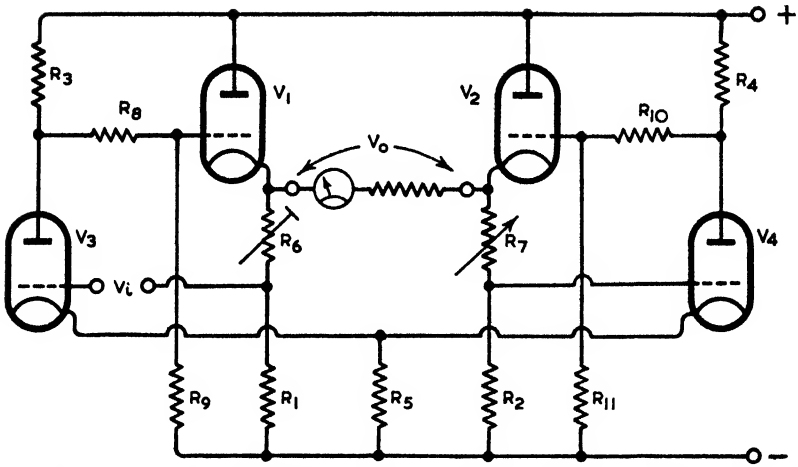


Fig 5.21.—Theoretical diagram of circuit for providing any ordinary voltmeter with virtually infinite input resistance without affecting calibration

(4) No significant change in calibration or zero setting on varying mains voltage 10% or output-valve characteristics 100%, without any supply-voltage stabilization.

A full account of this type appears in *W.W.*, Jan. 1952, and a briefer one in *Electronics*, Dec. 1951. Fig. 5.21 is the basic circuit. Each half of the amplifier is in effect a voltage stabilizer of the Fig. 4.6 type; the two halves are coupled by R_5. When a voltage V_i is applied to the terminals so marked, an equal voltage V_o appears between the output terminals. Exact equality is obtained by pre-setting R_6 and R_7, which compensate for the slight voltage loss in the amplifier; R_7 serves also for zero adjustment. Fig. 5.22 is the circuit of a high-grade "electronic test meter" embodying this system, which, with the values shown and lower-limit supply voltage, handles rather more than ± 50 V with practically perfect linearity. It was described fully in *W.W.*, Aug. 1954. Using a 0 to 3 mA meter, ranges with full-scale readings of $\pm 1{\cdot}5$, 5, 15 and 50 V are obtained by the usual multipliers in series with the meter; the higher ranges (150, 500 and 1,500 V) by a high-resistance input-potential divider. (This also forms part of the rectifier output circuit for a.v.; see Sec. 5.21.)

5.18. VALVE OHMMETERS

A valuable feature of a d.v. valve voltmeter is the ease with which wide-range ohmmeter facilities can be incorporated. Fig. 5.22 shows provision for mid-scale resistance ranges of 1, 10, 100 and 1,000 kΩ, applicable to the scale of Fig. 5.7, giving satisfactory reading from 100 Ω to 10 MΩ and rough readings over an additional decade at

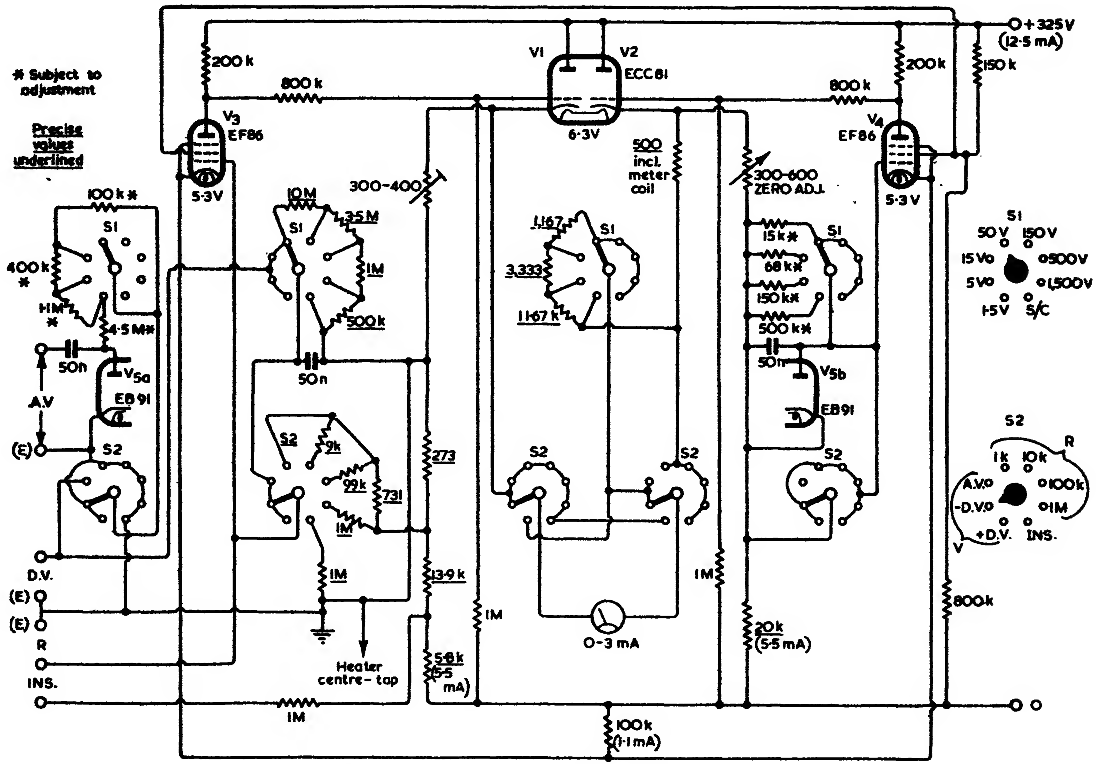


Fig. 5.22—Practical development of Fig. 5.18 for measuring direct and alternating voltages and resistance over a wide range. S1 and S2 each comprises a single ganged switch, with settings as shown on the right

each end. The resistance to be measured is connected between the terminals marked "R". Use is made of the very high input resistance of V_3 to include a facility for testing insulation, connected between the terminals marked "Ins.", with "(E)" as the guard terminal (Sec. 9.7). A special scale is needed for this, reading up to 1,000 MΩ, with indications up to 5,000 MΩ.

The Southern Instruments M1501 E.R.I.C. Universal Meter is a versatile d.c. instrument basically similar to that just described, with voltage, current and resistance ranges, and a "memory" device for storing voltage readings. The input current is only 10^{-14} A.

If very high resistances are to be measured it is generally better to design a special instrument for the purpose. Obviously the choice of input valve is of first importance. The author has found a selected and under-run EF37A (the ME1400 is virtually this) satisfactory for seven decade ranges from 0·1 MΩ to at least 1 MMΩ with rough indications up to 10 MMΩ (10^{13} Ω). The instrument is, in effect, a 0 to 5 V d.v. voltmeter; its main features are shown in Fig. 5.23. Standard test voltage for insulation is 500, but resistance usually depends very considerably on voltage, and a lower test voltage is provided to check voltage coefficient and for resistances that are unstable at the higher voltage. Stabilization of the test voltage is of the utmost importance, especially in measuring the leakage of capacitors. So, too, is the insulation of the input grid and the range switch—a ceramic type. Details of this instrument and its use have been given (*W.W.*, Nov. 1953).

The design and use of electronic megohmmeters are discussed by H. G. M. Spratt (*W.W.*, Oct. 1948), and details have been given of a General Radio instrument reading 0·5 MΩ to 2 MMΩ (*G.R.E.* Nov. 1951). For a simple low-reading megohmmeter (5 kΩ to 5 MΩ), see W. H. Cazaly (*W.W.*, Sept. 1949).

5.19. D.C. AMPLIFIERS

The design of direct-coupled amplifiers functioning essentially as resistance transformers rather than amplifiers has been considered in Sec. 5.16. Now we come to types giving a large voltage and/or current amplification, with the primary object of enabling smaller voltages or currents to be measured. If the ideal of infinite or zero input resistance respectively can be approximated, so much the better, of course. But the massive problem in high-gain d.c. amplifiers is stability—absence of zero drift. (Gain stability can be achieved, as with a.c. amplifiers, by use of negative feedback.) D.C. amplification has developed into such a large and complex subject that there is room for only a brief outline.

Where a high input resistance (many megohms) is necessary, valves are indicated. Transistors are more suitable where low input resistance is required (in current amplifiers) or can be tolerated. Although they are inherently more subject to internal variations than valves, especially

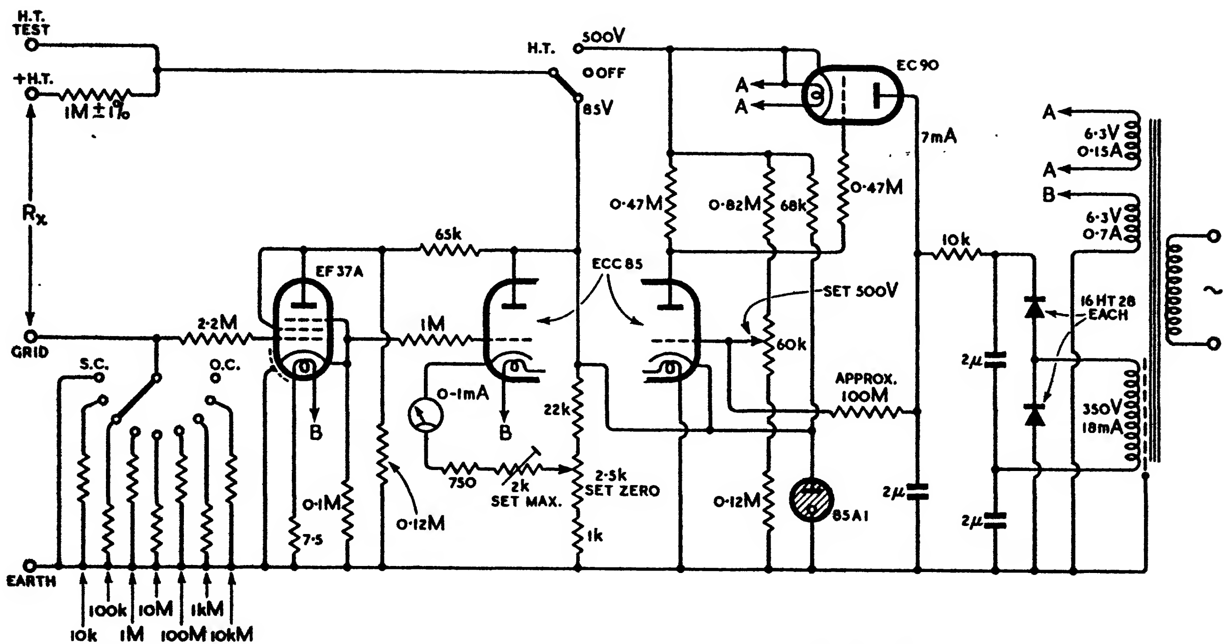


Fig. 5.23—Circuit of valve megohmmeter for measuring 0·1 to 1,000,000 MΩ at 500 or 85V

temperature effects, the absence of heaters and the generally better ratio of signal to feed current are a help, and transistor d.c. amplifiers can be produced with as good or better stability than valve amplifiers.

Apart from the question of valves or transistors, there are two main classes of d.c. amplifier: those in which there is direct coupling throughout, and various artifices are employed to minimize drift; and those in which the problem is dodged by "chopping" the d.c. signal into a.c. of a convenient frequency for amplification, and then restoring it to d.c. for the meter. This scheme is basically similar to the superheterodyne, and the "chopper" can be considered as a modulator or frequency changer.

The latter class can be subdivided according to the type of chopper employed: mechanical, electronic, photoelectric, Hall-effect, or capacitance modulator. The first of these is the most straightforward, but not quite so ideally simple as might be supposed. One's first attempt is likely to produce a substantial signal output without any input; this comes from thermoelectric contact effects and/or stray pick up, especially from the vibrator energizing circuit. Gold or platinum contacts and careful screening are essential. Nor is drift entirely avoided; slight changes in the setting of the contacts, so that the mark/space ratio varies, are liable to affect both the zero and the overall gain.

Nevertheless, this solution has been adopted in a number of commercial instruments. A good example is the Philips Type GM.6010 d.c. millivoltmeter. The d.v. to be measured is passed through a filter to remove any a.v. components, and is then converted to a square wave by a vibrator driven by a valve oscillator. The a.v. is amplified in a three-stage valve amplifier and fed to the tuned primary of an output transformer. Rectification between the secondary and the meter is effected by another contact on the vibrator. To ensure that the gain is correct for the calibration, it can be checked and adjusted by comparing the filament battery voltage direct with the same voltage after having been passed through an attenuator and the amplifier. This instrument is battery operated, has seven ranges from 1 mV to 300 V f.s.d., and the maximum error is stated to be less than 5%. The same firm has a 15-range instrument, GM.6020 (0·1 mV to 1,000 V f.s.d.). Another instrument using a mechanical chopper, intended as a substitute for a sensitive galvanometer in d.c. bridge work, and therefore confined to low ranges (0·3, 3 and 30 mV), also battery operated, is made by Rivlin Instruments. An obvious advantage, as compared with the galvanometer, is its relative immunity from mechanical and electrical damage, for a robust type of meter is used, and the current through it is limited by the amplifier. The Airmec Type N.855 is a d.c. amplifier only, for general use with meters, relays, etc. The chopper is a pair of relays energized by the 50 c/s supply, the overall voltage gain is normally 1,000, the zero drift is given as $< \pm 50\ \mu$V in 10 hours, and the input impedance is 4 MΩ and output 35–70 Ω. The Solartron AA.900 is a more elaborate d.c. amplifier with multiple feedback loops to achieve exceptional stability. The signal frequency range extends from zero to 40 kc/s (unusual for a chopper type), the voltage gain is variable up to 2,000 with an accuracy of $\pm$ 1%, the drift $<$ $\pm$ 2 μV over 40 hours, and input and output impedances 100 kΩ and $<$ 1 Ω respectively.

The idea of substituting semiconductor devices for mechanical contacts is attractive, but suffers from the disadvantage that neither open-circuit nor closed circuit is complete. Design is treated by H. Kemhadjian (*M.T.C.*, Dec. 1958).

A photoelectric chopper is employed in the Hewlett-Packard Model 412A d.c. valve voltmeter, which has 13 ranges from 1 mV to 1,000 V f.s.d., for which an accuracy of $\pm$ 1% of f.s.d. is claimed. The instrument also has 13 current ranges and 9 resistance ranges, and can be used as a $\times$ 1,000 amplifier. Modulation and synchronous demodulation, at input and output of the selective a.c. amplifier, is obtained by 50 c/s interruption of a light beam focused on photoconductors. Model 425A, working on the same principle, is even more sensitive: the lowest ranges are $\pm$ 10 μV and $\pm$ 10 $\mu\mu$A f.s.d.

Where an ultra-high input resistance is required (up to as much as 10^{15} Ω) the capacitance modulator is the choice. It usually takes the form of a vibrating reed, which periodically varies the capacitance in series with such a high resistance that the charge due to the d.v. to be measured remains practically constant throughout the vibrator cycle. Since $V = Q/C$, an alternating voltage is generated across the reed capacitance, and is amplified. The input resistance is limited only by the quality of the insulation. Because of this, such instruments are usually classed as electrometers. Examples are made by Electronic Instruments, Ericsson Telephones and Ekco Electronics.

In direct-coupled or " straight " d.c. amplifiers a favourite zero-stabilizing device is the connecting of valves or transistors in balanced pairs, either as in Fig. 5.20 or in " long-tailed pair " (i.e., with a common cathode resistor of relatively high value). Stabilization of feed voltages is another obvious means. So far as transistors are concerned, the article by Kemhadjian mentioned above is an excellent review of both straight and chopper design, with a comparative table of typical performance. Most of the trouble with transistors is due to temperature changes, affecting three parameters: (1) inverse leakage current I_{co}; (2) base-emitter voltage V_{be}; (3) amplification factor α_e. Unfortunately all three tend to increase collector current with temperature. (1) can be largely excluded, and (3) considerably reduced, by using silicon. (2) is minimized by a high external base resistance. The effects of short-term temperature changes can be further reduced by mounting the transistors in a substantial metal block, and still more by automatic temperature control. The tabulated comparison gives, as typical temperature coefficients of zero drift, 100 μV/°C for germanium or silicon long-tailed-pair amplifiers with a voltage gain of 500, improved to 5 μV/°C by temperature control. As little as 0·5 mμV/°C is given for a transistor chopper type with a gain of 10,000.

Though less subject to temperature effects, valves are more subject to feed fluctuations, and stabilization of heater voltage as well as h.t. is normally necessary. There are, however, methods of automatically applying drift correction, either by rapid periodical withdrawal of the amplifier from service, or continuous comparison of two amplifiers. Such methods are discussed by D. J. R. Martin (*E. & R.E.*, Jan. and Feb. 1958). Zero stability to within 100 μV is claimed without heater supply stabilization. Similar methods have been used very successfully with transistors. The same writer (*E. & R.E.*, Dec. 1957) describes a method of balancing valves so that heater-supply effects cancel out more exactly. And D. Allenden (*E.E.*, Jan. 1958) shows how zero

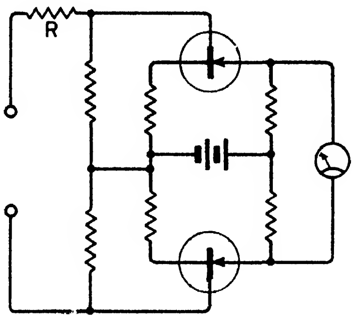


Fig. 5.24—Simple transistor d.c. amplifier enabling full-scale deflection of a multi-range moving-coil voltmeter to be obtained with 1μA input

drift can be countered, making use of an electrometer valve (ME 1403) having a filament current of only 8·2 mA.

It should be noted that there is usually little difficulty in making a straight d.c. amplifier work over a wide range of frequency from zero upwards, but use of the chopper technique leads to complications in this respect.

Fig. 5.24 is the simple basis of the circuit used in a pioneer transistor-amplifier voltmeter—the B.P.L. " Trans-Ranger "—in which *R* is 1 MΩ per volt full-scale. The ranges are from 0·1 to 500 f.s.d. The current drain is so small (about 0·5 mA) that the battery is permanently connected—there is no on/off switch. Because of the variableness of current amplification, provision is made for checking and adjusting the calibration.

Constructional details of a more sensitive instrument, intended as a null detector (d.c. galvanometer), are given by F. Oakes and E. W. Lawson (*W.W.*, Dec. 1957). It comprises two push-pull transistor stages: long-tailed-pair first stage, and 50–0–50 μA meter connected between emitters in the second stage. A fifth transistor is used for temperature compensation of the first stage. The transistors are housed in a thermally insulated aluminium block. The lowest range is 0·05 μA f.s.d., and short-term zero drift is not more than 5% of f.s.d. in 10 minutes. A somewhat similar instrument is marketed by Ferguson Radio. The power supply is a mercury battery, having a life of about six months continuous running.

5.20. SIMPLE A.C. VALVE VOLTMETERS

Before going on to combinations of high-gain d.c. amplifiers with a rectifier for measuring a.v., brief note should be taken of three single-valve types, the basic circuits of which are shown in Fig. 5.25.

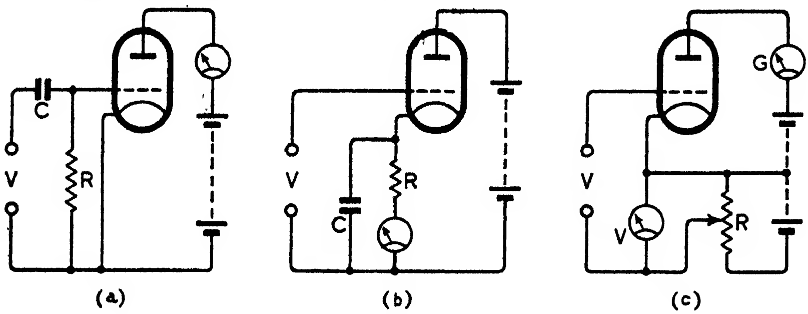


Fig. 5.25—Basic circuits of three types of a.v. voltmeter in each of which a single valve is used as d.c. amplifier

(1) *Cumulative-Grid.* This is just the old leaky-grid detector, which can be regarded as a Fig. 5.15 diode rectifier (grid and cathode) incorporated in a single-stage d.c. amplifier. It is the more sensitive of the two original Moullin valve voltmeters, and because maximum deflection of the meter is with zero input the pointer cannot be deflected off the scale. For the same reason, zero stability is poor and the scale is non-linear. In the Marconi Instruments TF.899A valve millivoltmeter (Fig. 5.26) the natural sensitivity of this type is increased, and the zero stabilized, by using it in a bridge circuit balanced by a second valve. In this way a bottom range of 20 to 150 mV is obtained without pre-detector amplification. The instrument, which has a probe, is effective from 50 c/s to 100 Mc/s.

(2) *Reflex.* In this type, a Fig. 5.13 diode rectifier is incorporated in a cathode follower. It is, in fact, sometimes called a cathode follower or auto-bias valve voltmeter. The d.v. negative feedback makes it a more stable and less sensitive type than (a). It lends itself to multi-ranging by varying *R* in steps, but the supply voltage must be adequate —of the order of three times the highest peak signal voltage. The scale is almost perfectly linear above about 2V. High readings are approximately proportional to peak values, but low readings are something between peak and mean. This rather indefinite relationship is a disadvantage if voltages of various waveforms have to be measured. Omitting C converts it into a true cathode follower and multiplies the scale readings by about 3; at the same time the readings are fairly accurately proportional to mean values of positive half-cycles. This

modification is best restricted to a.f., owing to the effect of stray capacitance across R.

The accuracy of this type is vulnerable to supply-voltage variations; but in an article analysing it (*W.W.*, Oct. 1949) the author has explained a simple modification to overcome this disadvantage—a resistance between anode and meter, equal to μR. The theory is, in fact, based on the μ bridge (Fig. 7.22). Fig. 5.27 shows a practical three-range circuit using a 0 to 1 mA meter, incorporating this modification which

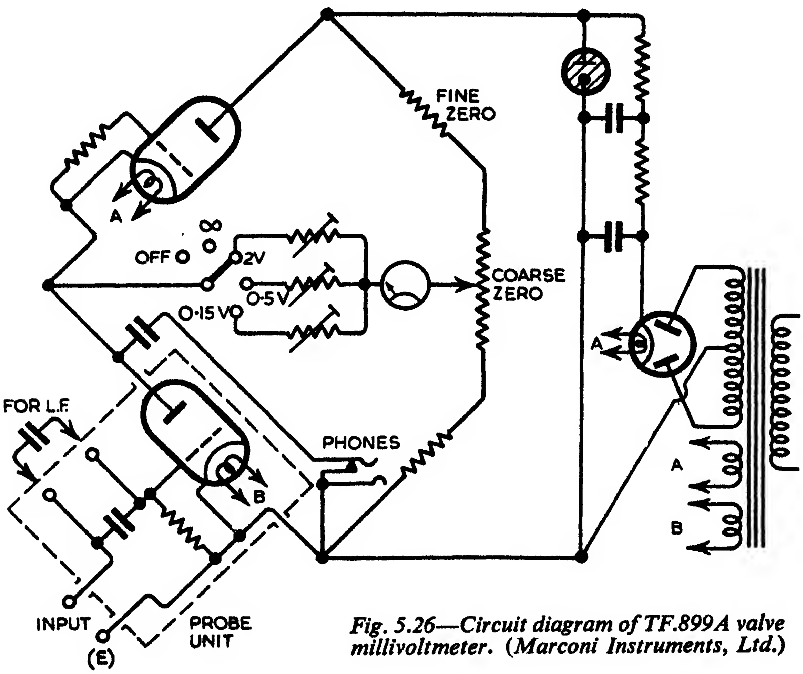


Fig. 5.26—Circuit diagram of TF.899A valve millivoltmeter. (Marconi Instruments, Ltd.)

renders supply-voltage stabilization unnecessary. In Fig. 5.28 the curves without the modification (full line) and with (dotted) show how effective it is, not only at zero input but at all signal levels. For high frequencies the electrolytic capacitors earthing cathode and anode should be supplemented by r.f. types of, say, 5 nF, connected close up. The 5 MΩ resistor is to prevent the grid from being open-circuited, and if desired a push-switch can be fitted to open-circuit it when working on high-impedance circuits.

(3) *Slide-Back*. In this type, R is first adjusted so that with no V the anode current is reduced to zero, or rather (because the " zero " is indefinite) some very small current, indicated by G. When V is applied, R is readjusted to give the same small reading on G; the

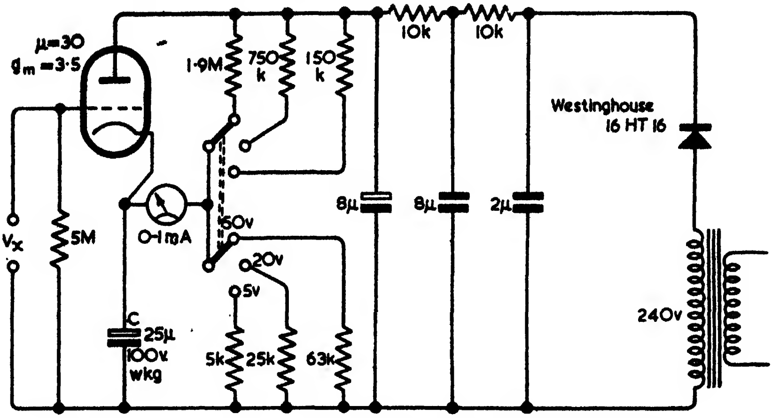


Fig. 5.27—Circuit diagram of reflex valve voltmeter with supply-voltage compensation. It is a useful and easily-made type for general purposes

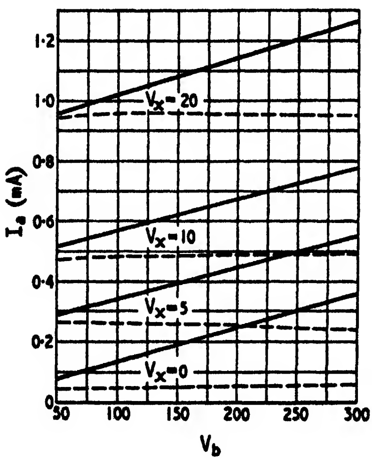


Fig. 5.28—The full lines, which apply to a reflex valve voltmeter without compensation, show how the meter deflection depends on the supply voltage, V_b. The readings were taken with $R=55$ kΩ, $\mu=30$, g_m (nominal)$=3{\cdot}5$ mA/V, and values of 50-c/s sinusoidal voltage. The dotted lines show the effect of adding compensating resistance (=μ times load resistance) from anode to load resistance

increase in negative bias, read on the voltmeter, is ideally equal to the peak value of V. In practice there is an appreciable discrepancy, even if the " zero " current is less than 1 μA. So, as it calls for a sensitive current indicator as well as a voltmeter, and an adjustment is needed for getting a reading, the slide-back type is unpopular in spite of its advantages—almost complete absence of error due to variations in supply voltages and valve characteristics, measurement of peak values using ordinary d.c. voltmeter calibration, ease of discriminating between nearly equal large voltages or setting a voltage to a given level, and low input loss.

5.21. DIODE-PLUS-AMPLIFIER VOLTMETERS

What can almost be called the standard commercial type of valve voltmeter comprises a d.v. valve voltmeter and a diode rectifier which can be tacked on at the front for a.v. ranges. Because of the loss of rectification efficiency below about 0·3 V, measurement much below that level is seldom attempted, so little or no voltage amplification is required. The " amplifier " is therefore more correctly considered as a resistance transformer (Sec. 5.16) for giving a low-resistance meter a nearly infinite input resistance. Because of the characteristics of the diode rectifier, discussed in Sec. 5.13, this high resistance is needed not only for d.v. measurements but for reasonably accurate a.v. measurements. Since stability is needed rather than voltage gain, the amplifier usually consists of a twin balanced cathode follower, with either a single valve on each side or (for better performance) two, as for example the circuit described in Sec. 5.17.

The circuit diagram, Fig. 5.22, shows how the balanced-valve technique is often extended to the a.v. portion, the working diode having its twin counterpart on the other side. This is even more necessary for a.v. than for d.v., because with a thermionic diode the zero is displaced to an extent that depends on the cathode temperature. In the example shown, the d.v. input potential divider is augmented for a.v. to just the extent needed (very nearly $\times \sqrt{2}$) to reduce the rectified peak values to r.m.s. Above 0·5 V, a.v. can be read off the linear d.v. scale. The 500 V and 1,500 V ranges are omitted for a.v. because of the limited voltage rating of diodes suitable for the low ranges. Television e.h.t. diodes such as the EY51 can be used for high voltages, but are limited in frequency by transit-time effect.

Commercial instruments of this general type are too numerous to list in full. The author's design shown in Fig. 5.22 is the basis for the Jason EM10, available as a kit or ready-made. The typical specification shows a number of ranges from 1 V f.s.d. upwards, an input impedance of 2–7 pF in parallel with a few megohms, and a top frequency limit of 100–300 Mc/s, besides d.v. and resistance ranges. The Airmec Type 217 and Marconi Instruments TF1300 come into this class. The latter firm's TF1041B (Fig. 5.29), which uses a circuit akin to Fig. 5.22, is exceptional in frequency coverage (20 c/s to 1,500 Mc/s) because it uses a disk-seal diode. It should be understood that input resistance is of the megohm order only up to a few Mc/s; it falls off at an increasing rate and is only in kilohms at the top frequency. The fact that v.h.f. circuits are usually measured in a resonant state and are comparatively low in impedance is a consoling thought. An input probe is standard practice. See also Sec. 12.9.

An exception to the rule that diode-first voltmeters cannot measure much below 50 mV is the Boonton Model 91–CA, in which a germanium diode is followed by a chopper-type d.c. amplifier. It has eight ranges, from 1 mV to 3 V f.s.d., and measures down to 0·3 mV. Maximum error is given as 5% of f.s.d. at 200 Mc/s and 10% at 600 Mc/s. Because it works mostly on the nearly parabolic part of the diode characteristic, readings are r.m.s. below 0·1 V. The input resistance is inevitably low, but the capacitance is given as 2·5 pF. In the Airmec Type 301, which has similar voltage and frequency ranges, plus d.v. ranges, a varistor (non-linear resistor) is used to linearize the a.v. scales. The Hewlett-Packard Model 411A uses an ingenious feed-back comparison device to obtain scale linearity.

5.22. AMPLIFIER-PLUS-DIODE METERS

The ability to measure signal amplitudes of the low millivolt order on a linear scale is very valuable, and the usual means is pre-rectifier

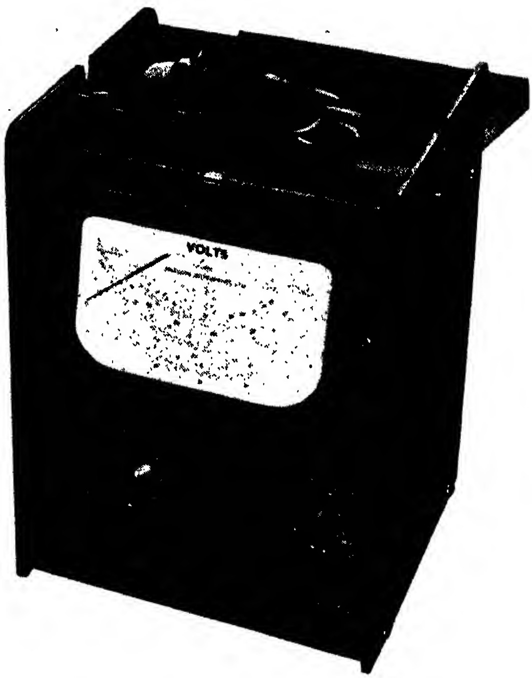


Fig. 5.29—Wide-frequency-range valve voltmeter (TF1041B) using a balanced d.c. amplifier. The probe, stored in the top compartment, contains a disk-seal diode, effective up to 1,500 Mc/s. (Marconi Instruments, Ltd.)

or signal-frequency amplification. The problem is to obtain adequate and stable amplification over the signal-frequency range one wants. In general, amplifier gain and bandwidth are inversely proportional to one another.

A design in which these conflicting requirements are met to an exceptional extent is the Philips GM6014 millivoltmeter. It has 10 ranges, from 1 mV to 30 V f.s.d., and reads down to 100 μV. The frequency band is 1 kc/s to 30 Mc/s. Accuracy is better than 2·5% of f.s.d., added to a frequency error of up to 5%. Seven stages of amplification are needed. The first is a cathode follower, located in the probe, to give a high input impedance—7 pF in parallel with 3 MΩ at 1 kc/s, down to 50 kΩ at 30 Mc/s. Its cathode resistance is the range attenuator. Negative feedback and stabilization of h.t. and heater voltages are used to stabilize and linearize the amplification, and a stabilized 30 kc/s *RC* oscillator is incorporated for checking it at 30 mV and 3 V (± 1%). The output is rectified by crystal diodes and indicated by a moving-coil meter. For ranges above 0·3 V a capacitive attenuator is pushed on to the probe, raising the lower frequency limit to 10 kc/s and increasing the input impedance to 2 pF and 2 MΩ at 30 Mc/s. The basis for design was explained in *Philips Technical Review*, Jan. 1950. A comparable specification, but with a lower frequency limit (30 c/s), applies to the Marconi Instruments TF1371; see L. G. White (*M.I.*, Sept. 1960). Its amplifier valves have their g_m stabilized by z.f. negative feedback, provided by high-valued cathode resistors.

Where high accuracy rather than high frequency is required, it can be found in the Solartron VF252 millivoltmeter, in which the error is given as ± 1% of f.s.d. from 15 c/s to 100 kc/s. There are nine ranges, from 1·5 mV to 15 V, and an attenuator to extend these to 150 V. The amplifier comprises four " ring-of-three " (two pentode stages and cathode-follower output) sub-amplifiers, each with negative feedback; and the h.t. is stabilized 400 : 1. A special feature is a 2·5 V output for an oscilloscope.

Mention of oscilloscopes, which are very soon to be considered, is a reminder that this same problem of signal-frequency amplification is one that has been very intensively studied in connection with those instruments, and extension of a.v. millivoltmeters up to 100 Mc/s or more may be achieved by similar means; namely, distributed amplifiers (Sec. 5.30).

For a good review of modern a.v. voltmeters, more especially the pre-amplifier kind, see *E.T.*, March 1960.

Transistors are in general not so suitable as valves for this kind of amplifier, but a design by R. R. Vierhout (*E.E.*, July 1960; see also Oct. 1960, p. 644) is interesting for having an input resistance of 1–6 MΩ, thanks to negative feedback. Its frequency range is 1 c/s to 100 kc/s, and the lowest range 10 mV f.s.d.

Work on transistor circuits creates a demand for signal-current meters. A.c. microammeters by Dawe (Type 618) and Quan-Tech Laboratories (Model 301) are somewhat similar in specification. Both employ transistor amplifiers and provide alternative input connections: a clip-around probe (Sec. 5.11) for moderate sensitivity without breaking circuit, and an insertion probe for high sensitivity. Both have an output connection for oscilloscope, quote ± 2% accuracy for insertion connection and ± 5% for clip-around, have a top frequency limit of 100 kc/s, and can be worked from the mains; but the Dawe instrument is alternatively battery-driven. The ranges, using the clip probe are (Dawe) 3 mA to 1 A, (Q.T.L.) 0·3 mA to 0·1 A; and using insertion (Dawe) across 0·3 Ω, same ranges, or, across 300 Ω, 3 μA to 1 mA;

(Q.T.L.) 3 μA to 0·1 A using a current transformer. An alternative transformer gives ranges of 0·3 mA to 10 A. The Q.T.L. amplifier has six stages, using transistors in pairs, each pair with negative current feedback. Direct coupling is used throughout.

5.23. SELECTIVE METERS

There are several types of instrument that can be regarded as special valve voltmeters (though not usually so named), in which the amplifier is more or less selective. Some examples now follow.

(1) *Null Indicators.* An indicator to enable a signal to be adjusted to zero is a necessity in all null methods (Secs. 3.10 and 7.3). A galvanometer for d.c. and phones for a.c. are the classical indicators, but they have obvious limitations. Sensitive galvanometers are mechanically and electrically fragile; electronic substitutes are included in Sec. 5.19. Although phones are remarkably sensitive they are ineffective outside the middle audio frequencies, or in noisy situations, or where the signal to be nullified is not alone. What is wanted is an instrument that shows whether the signal is increasing or decreasing, over a very wide range, and especially near zero; and preferably tunable to the signal frequency. A visual indicator such as a meter or " magic eye " (Sec. 5.42) is desirable where there is much noise, with provision for loudspeaker and phones where there is not. Manual gain control is a nuisance, and the shock of accidentally going right off balance when set to maximum gain is relished neither by instrument nor operator. Some sort of a.g.c. is desirable, but with a rather longer time constant than usual so that bridge adjustments cause a clear transient up or down swing to show which way to turn next for balance.

Two of the author's designs, one with a milliammeter and the other with a " magic eye ", have been described (*W.W.*, May 1951). Both are mains-driven, and can be either " flat " or tuned sharply over bands of frequencies around 50, 400, 1,000, and 2,000 c/s by means of a feedback circuit, similar to that in Fig. 4.29, set just short of oscillation. The effective range of signal is 10 μV to 10 V, without manual adjustment.

(2) *Wave Analysers.* What is called a wave analyser is a highly-selective high-gain valve voltmeter, and is a powerful aid to a.f. investigation, enabling one to measure the individual harmonics and intermodulation products of non-linearity distortion; to analyse noise, hum, vibration, etc.; and to measure the frequency characteristics of systems fed with " white noise ". There are two main types of wave analyser, analogous to t.r.f. and superhet receivers. In the first, selectivity is usually obtained by *RC* frequency-discriminating feedback,* such as that used in the null indicator just mentioned. The Muirhead-Pametrada wave analyser (*Muirhead Technique*, April 1950) is of this type, which normally has a bandwidth proportional to the

* " Some Applications of Negative Feedback with particular reference to Laboratory Equipment ", by F. E. Terman, R. R. Buss, W. R. Hewlett, and F. C. Cahill. *Proc. I.R.E.*, Oct. 1939.

frequency tuned in. It is as if the tuned circuit had the same Q at all frequencies.

In the other type the incoming signal is heterodyned to shift its frequency to that of a fixed-tuned filter, usually about 50 kc/s. The result is that the bandwidth is the same at all frequencies. In the General Radio 736-A model it is only 4 c/s, which represents much higher selectivity at the upper audio frequencies than is obtained with the " straight " type. The heterodyne type is therefore the choice for analysing intermodulation sidebands at these frequencies, but is

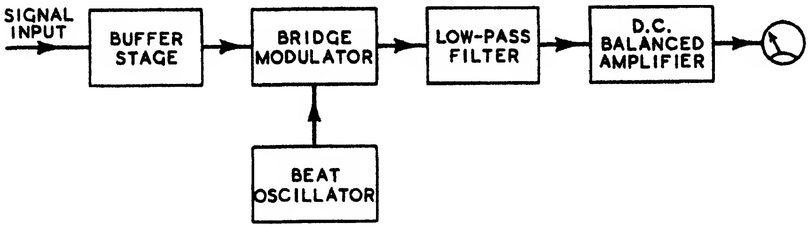


Fig. 5.30—Block diagram of author's zero-i.f. wave analyser

unsuitable if the signal frequency is not very constant. This disadvantage is to some extent countered in the Hewlett-Packard Model 302A heterodyne wave analyser by an a.f.c. system which keeps the frequency difference between the heterodyne oscillator and the signal constant at the i.f. (100 kc/s) over a range of 100 c/s. Other interesting features are introduced in this model. It is completely transistorized—with the usual benefits, including mains or battery drive; the frequency range extends from 20 c/s to 50 kc/s; there is an oscillator tracking with the analyser voltmeter tuning over this frequency range, for use as a source for measuring frequency/amplitude characteristics of filters, etc.; and provision is made for the frequency of the measured signal component to be restored so that it can be measured precisely by an electronic counter. Its design is discussed in *Hewlett-Packard Journal*, Sept.-Oct. 1959.

Apparatus of this kind is of course expensive, and is beyond the amateur to make. To by-pass the difficulties of a 50 or 100 kc/s tuned amplifier with a bandwidth of only 3 or 4 c/s, the author adopted zero frequency as the i.f., since all that is required then is a simple *RC* low-pass filter (Fig. 5.30). No meter rectifier is needed either, the output frequency being low enough for the peak-to-peak amplitude to be measured with a moving-coil milliammeter in a stable d.c. amplifier of the Fig. 5.21 type. When the frequency of the beat oscillator is set a fraction of 1 c/s different from that of the signal component being measured, the phase difference between the two slowly alters throughout 360°, giving successive positive and negative deflections. Details have been given (*W.W.*, Aug. 1955, where R_7 in Fig. 2 should be 75 kΩ, not 33 kΩ).

The down-to-zero i.f. idea is adopted also in the Airmec r.f. wave analysers: Type 853 for 30 c/s to 30 Mc/s, and Type 248 for 5 Mc/s to 300 Mc/s. Commercial wave analysers have been reviewed by R. Brown (*B.C. & E.*, Oct. 1960).

(3) *Phase-Sensitive Voltmeters.* The Fig. 5.30 wave analyser can be regarded as a phase-sensitive voltmeter, for the deflection at any instant is proportional to the peak amplitude of the signal multiplied by the cosine of the phase angle between it and the local oscillation—which acts as a reference signal. In phase-sensitive voltmeters designed as such, provision is made for measuring both in-phase and quadrature components. The instrument described by R. Kitai (*E. & R.E.*, April 1957) is broadly similar to Fig. 5.30 with the addition of a 90° phase shifter that can be switched into the oscillator (reference signal) channel. The modulator (called a coherent detector) is a bridge containing two thermionic diode rectifiers, compared with four selenium diodes in the wave analyser. The frequency band is 20 c/s to 40 c/s.

In the Solartron VP 250 "Resolved Components Indicator", in-phase and quadrature components can be read simultaneously on two indicators, which are of the wattmeter type. The amplified signal is fed to both; the reference signal is fed directly to one and via a 90° phase shifter to the other. To make the instrument effective from 20 c/s to 20 kc/s, no transformers are used and the " wattmeters " are of a special thermocouple type.

(4) *Loudness and Hum Meters.* For measuring the intensity of sounds as judged by ear, i.e., their loudness, it is necessary for the a.f. amplifier to have a frequency characteristic similar to that of the ear (Fig. 14.37). An awkward complication is that this frequency characteristic varies with intensity, so sound meters which have to cover a wide range of intensity require some means of varying the characteristic, either continuously or in steps. For measuring hum or other sounds presumed to be at a level not very far above the threshold of hearing, and where the differences in sensitivity of the ear at different frequencies are particularly large, one fixed characteristic is enough. Fig. 5.31 is a circuit of an amplifier " weighted " to compensate for the characteristic of the average ear at about the 20-phon level of loudness, which may be regarded as typical of unobjectionable hum.

Fig. 14.37 shows that the slope between 20 and 800 c/s is about 12 dB per octave. A slope of 6 dB per octave can be obtained by taking the potentials developed across an inductance in a constant-current circuit, such as pentode whose internal resistance is very much greater than the maximum load impedance. The required characteristic can therefore be obtained by employing two stages in cascade. The inductance is selected by considering the frequency at which no gain is necessary. Assuming a pentode with a working mutual conductance of 2 mA/V, an inductance of 1¼ H gives a small gain at 100 c/s, which is the lowest main component in a full-wave rectifier system, and a slight loss at 50 c/s, which is prominent in a half-wave system. The resistance of the coil should not exceed about 200 Ω at the most.

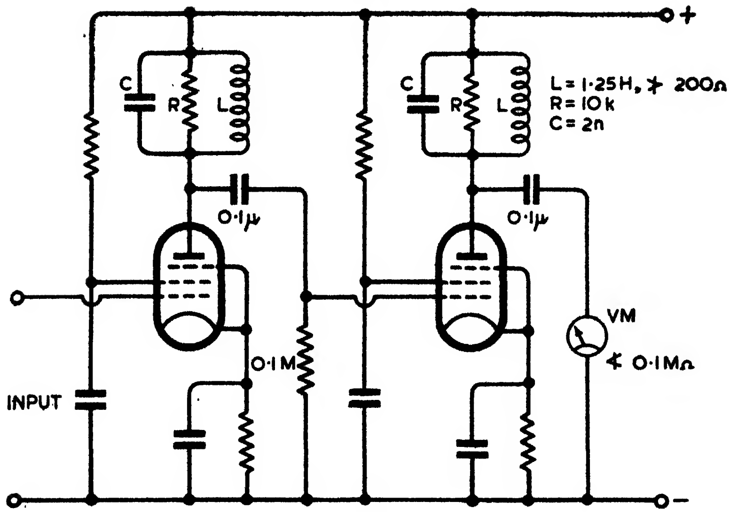


Fig. 5.31—Circuit of a " weighted " amplifier to compensate for the characteristics of the average ear (Fig. 14.37) to indicate loudness of hum or other low-level sound

To limit the gain at the upper frequencies and bring the compensation into accord with the remainder of the aural characteristic, the coil is shunted by 10 kΩ resistance and 2 nF capacitance. The approximate overall gain is:

f: c/s	30	100	300	1,000	3,000	10,000
Gain: dB	−10	8	28	42	48	42

and the resultant of this and the 20-phon aural characteristic is almost horizontal. The output voltmeter should either have an impedance not less than about 0·1 MΩ, or be driven from a cathode-follower stage, and is preferably of the r.m.s. variety.

Additional complexities enter into the measurement of noises, some of which are continuous and others impulsive. It is necessary to adjust the time-constant and waveform characteristics of the instrument to make the readings correspond with what is judged to be the relative loudnesses of the noises.*

5.24. RECORDING METERS

There is often a need to record meter readings continuously on paper. The result is primarily a graph of current, voltage, etc., against time.

* " Electrical Noise ", by D. Maurice, G. F. Newell, and J. G. Spencer. *W.E.*, Jan. 1950.

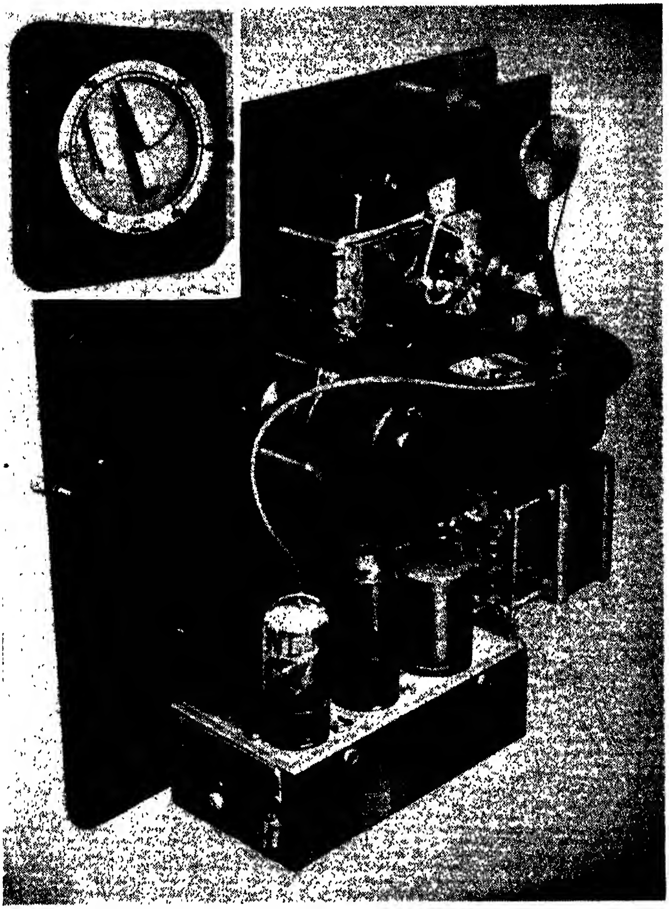


Fig. 5.32—Interior of Fielden " Servograph " recording meter, a front view of which is inset. (Fielden Electronics, Ltd.)

But if, say, the frequency of the signal causing the deflection recorded is made to vary in a known manner with respect to time, one can obtain a signal amplitude/frequency graph.

The most obvious way of making a recording meter is to substitute a very light friction-free pen or stylus for the usual pointer, and arrange it to sweep over a roll or disk of paper kept slowly moving by a clockwork or electric motor. The more accurate and sensitive the meter, the less likely is it to be relied upon to keep on writing plainly, so alternative writing methods obviating direct contact have been tried, such as marking the paper by sparks passing periodically through it from the tip of the pointer to a metal support. However, the design of conventional writing systems is now good enough to give satisfactory results, with simplicity, provided that the meter part can be allowed rather more power than is usual in electronic apparatus—say 1 mW or more. Where this is not allowable, the tendency is to adopt servo technique, whereby the power to move the pen is derived from a motor and the direction in which it moves is determined by the difference between the actual signal voltage, etc., and that indicated by the position of the pen. Plenty of power can thus be provided for moving a robust and reliable pen—and extra relays or controls, if needed.

An example of a servo-operated recording system of comparatively moderate cost, applicable to any of the usual types of meter move-

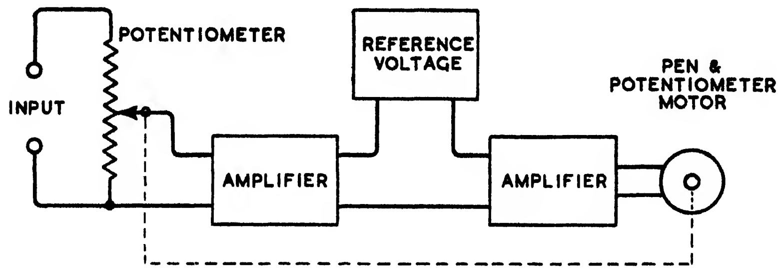


Fig. 5.33—Block diagram of potentiometer-type recording meter

ment—down to a few microwatts for f.s.d.—is the Fielden " Servograph ", Fig. 5.32. In place of the usual pointer is a light metal vane forming one plate of a variable capacitor. Moving in the same arc is a similar vane driven—along with the pen—by the servo mechanism. The spacing between the two is maintained constant by an electronic capacitance relay which controls the servo motor. The pen sweeps over a radius of a rotating disk form of paper chart.

Most other recording meters work on a potentiometer principle, in which the only moving parts are those driven by the servo motor. The pen, which runs along a straight track (thereby enabling conventional squared paper to be used, instead of special curved kinds), is combined with the moving contact of a potentiometer, across which the signal to be recorded is connected. The portion tapped off is compared—usually after amplification, and, if a.v., rectification—with a constant reference voltage. The difference between the two is greatly amplified and used to drive the motor in the direction which reduces the difference. Fig. 5.33 shows only the basic principle; in practice there are complications such as velocity feedback to prevent mechanical oscillation (" hunting ") of the movement.

Instruments of this kind intended for use in the communications field usually have facilities for taking amplitude/frequency characteristics; for instance, logarithmic potentiometer, and provision for coupling to the frequency control of a signal generator. An example is the General Radio Type 1521A, which is transistorized (*G.R.E.*, June

1959). Bruel & Kjaer have a wide range of co-ordinated equipment of this kind.

Some types of recording instrument are designed to be capable of accurately following the signals through them at much higher speeds than ordinary meters; if necessary, up to several thousand c/s. They are then usually known as oscillographs. Where permanent records are of primary importance—especially a number of simultaneous records of different quantities—and the frequencies are not impossibly high, such instruments are still much used. This brief reference must be sufficient as we press on to the vastly more important cathode-ray oscilloscope.

For further information on all types of electronically aided meters, see *Electronic Measuring Instruments* (2nd ed.), by E. H. W. Banner (Chapman & Hall, 1958).

5.25. CATHODE-RAY TUBES: ADVANTAGES

Although it is not usually considered as a voltmeter, the cathode-ray tube is often a practical alternative to the valve voltmeter. As such, and no more, it is perhaps not always the most convenient choice. But whereas meters show merely the *magnitude* of the voltage or other quantity measured, the cathode-ray tube is capable of showing (simultaneously, if necessary) the magnitude, form, phase, and frequency of a wave, and its relationship to other quantities. And that is only starting on its applications. In fact, it gives vastly more information and insight into the workings of electrical and allied apparatus than is possible by any other means. The X-ray tube is no more indispensable to the surgeon than the cathode-ray tube to technicians and scientists, as well as engineers.

An admirable feature of the c.r.t. is that if the strength of the signal applied is, say, a hundred times greater than was expected, it does not leave one ruefully surveying the ruins of one's best meter. This aspect may seem to be having excessive prominence here, but accidents do happen in the best-regulated laboratories; if by extreme care and attention they are avoided, such care and attention are diverted from the main business in hand.

And cathode-ray oscilloscopy is not necessarily expensive. While it is true that ready-made oscilloscopes can cost as much as £1,250 plus optional extras, others capable of valuable work can be obtained for about £30. And while it would be wrong to encourage any lab. worthy of the name to do without at least one complete and versatile oscilloscope, it should be noted that much can be done with a bare cathode-ray tube, no very high voltages, and even no time-base generator. The necessity for a time base is often assumed when in fact it is not necessary, or perhaps (as will appear in connection with the measurement of distortion) not even the most satisfactory technique. And the conventional setting of a c.r. tube, in an oscilloscope with much switching and wiring between the deflection-plate terminals on the panel and the plates themselves, is liable to introduce undesirable effects that can be avoided with a separate tube.

5.26. CHARACTERISTICS OF OSCILLOSCOPE C.R. TUBES

At least an elementary knowledge of the principles and construction of cathode-ray tubes is assumed.* The following is a summary of the characteristics of oscilloscope tubes as distinct from those used for television.

They all have two pairs of plates for electric deflection (Fig. 5.34), distinguished as the X and Y pairs (corresponding to X and Y axes of

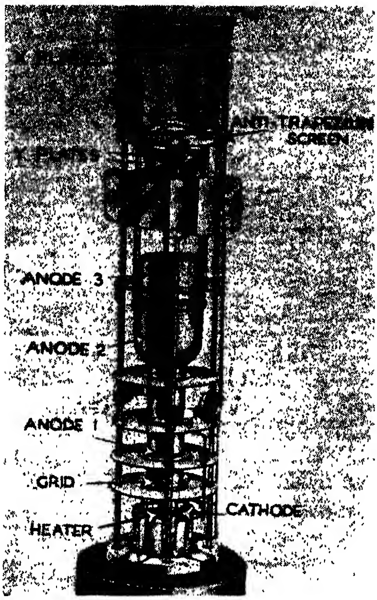


Fig. 5.34—Typical electrode construction of single-beam cathode-ray tube. (A. C. Cossor, Ltd.)

a graph, X being horizontal and Y vertical). But this does not exclude the use of magnetic deflection, alternatively or additionally, by means of external coils against the neck of the tube as in Fig. 5.35*a*. It is not so generally useful, because appreciable power is required to cause deflection, and the inductance of the coils is liable to create difficulty except at very low frequencies, but it is convenient for 50 c/s deflection. A pair of coils enables a nearly uniform deflecting field to be obtained, which is not so with one coil. Note the plane of deflection, at right angles to the coil axis, compared with electric deflection (*b*). Whereas electric deflection is inversely proportional to the anode voltage, magnetic deflection is inversely proportional to the square root

* They are introduced in the author's *Foundations of Wireless* (7th ed., Iliffe, 1958). A simple book on the subject is *The Cathode Ray Oscilloscope*, by Harley Carter (2nd ed., Cleaver-Hume, 1959). For more detailed treatment, with many references, see *The Cathode-ray Tube*, 3rd ed., by G. Parr and O. H. Davie (Chapman & Hall, 1959).

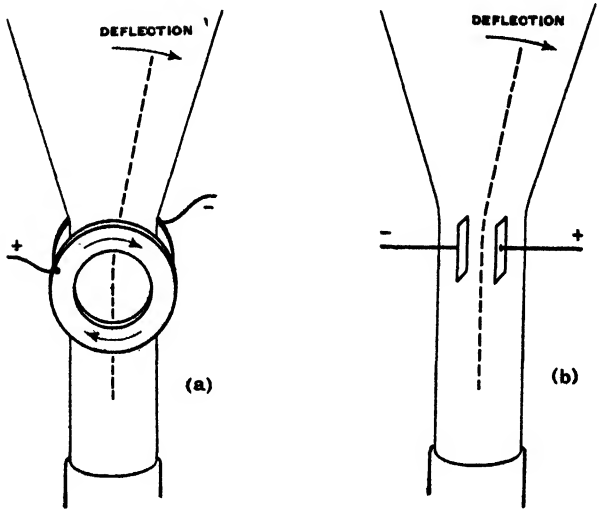


Fig. 5.35—Showing how (a) *magnetic deflection and* (b) *electric deflection are accomplished, and the relationship between polarity of current or voltage and the direction of deflection*

of anode voltage (Sec. 14.29) so is most worth considering when the voltage is very high.

Electric deflection sensitivity is usually stated in millimetres (measured on the face of the tube) per volt applied between a pair of plates,* and obviously depends on the length of the tube, but a typical value is $400/V$, where V is the voltage of the final anode.

The maximum V for which the tube is rated—usually several thousands—thus necessitates a considerable signal voltage to give a reasonable-sized deflection. And great care must be taken not to let the spot rest stationary, for it quickly burns it, leaving an insensitive patch.

On the other hand, a very low anode voltage fails to give a good focus, or sufficient brightness unless the spot is moving slowly. In fact, for photographing high-speed transients 10 kV or more may be needed. Tubes which may be used for such purposes, or in the better general-purpose oscilloscopes, are provided with what is called a post-deflection accelerator (p.d.a.) or intensifier. This is one or more extra anodes on the screen side of the deflection plates. The simplest consists of a conducting band around the inside of the tube near the screen, connected to a high positive voltage. A larger p.d.a. ratio can be obtained without distortion of the trace if the accelerating voltage is distributed over the distance between deflectors and screen. A usual means is a

* Alternatively it is sometimes given in an upside-down but perhaps more directly useful form as volts per cm.

potential divider of the order of 20 MΩ consisting of a carbon spiral around the conical part of the tube. Some tubes have more than one stage of p.d.a.

In the Ferranti Series 5/63 tubes the high sensitivity of 4 mm/V (2·5 V/cm) is obtained with 10 kV on a spiral p.d.a., the final pre-deflection anode voltage being only 1·2 kV. Besides that, the Y plates are long and close together, and there is what is called a beam lever, being a cylindrical electron diverging lens between X and Y plates, whereby the Y sensitivity can be controlled without affecting X deflection.

The voltage of the final pre-deflection anode, which determines the deflection sensitivity, is made no more than necessary for a satisfactory focus—usually from a half to a tenth of the p.d.a. voltage. The voltage V is therefore now seldom above 2 kV and is often as low as 1 kV. In simple miniature tubes, used mainly as monitors in valve circuits, it may be 400–800.

The usefulness of an oscilloscope pattern or trace depends not so much on its size as on the ratio of size to spot diameter, so fineness of focus is important. In general the focus improves with increasing V, which tends to compensate for the decreasing deflection sensitivity.

Tubes for oscilloscopes are provided with electric focusing, obtained by adjusting the voltage applied to one or sometimes two intermediate anodes. In the Emitron 1CP1—a miniature c.r.t. on a valve base—focusing is obtained automatically over the working range of anode voltage (500–800). In tubes with electric deflection and focusing, an asymmetrical deflecting voltage—one that changes the mean voltage of the plates by not balancing positive on one by equal negative on the other—upsets the potential pattern and hence both deflection sensitivity and focus, causing trapezium distortion and deflection defocusing respectively. Provision of a push-pull deflection amplifier to avoid these effects is not always convenient, but the plates of some tubes are designed to render it largely unnecessary—note, for example, the curved-slot screen between X and Y plates in Fig. 5.34, the theory of which is explained by B. C. Fleming-Williams (*W.E.*, Feb. 1940).

If the deflection sensitivity is, for example, $500/V$ mm per volt, and the final anode is run at 1 kV, about 50 V is needed between a pair of plates to deflect the spot 1 in. An alternating deflection voltage draws a line on the screen; its length is a measure of the peak-to-peak value, so if the voltage were sinusoidal a 1-in line would indicate $50/2\sqrt{2} = 17{\cdot}7$ V r.m.s. It is important to realize that except in a few tubes that make it a special feature the deflection sensitivities of X and Y plates are unequal. In very short tubes the sensitivity of the plates nearer the screen may be only about half that of the plates nearer the cathode. Since there is more likely to be ample voltage for the X deflection, derived from a time-base generator, the plates nearer the screen are designated the X plates, leaving the greater sensitivity for the Y or signal plates.

Obviously, the greater the length of tube, the greater the deflection sensitivity; but to gain the full advantage of a long tube it is necessary for the focusing to be of a correspondingly high accuracy. For accurate work, the tubes provided with flat screens are to be preferred to those with convex ones.

Various types of screen are available, some being specially suitable for visual work and others for photographic; usually the former give a green light and the latter blue. The time taken for the fluorescence of the screen to disappear after the ray has moved from it may be anything from a fraction of a microsecond to half a minute or more; it is thus possible to obtain special screens for work involving exceptionally fast or slow deflections. A modified type of tube called the Memotron* displays the trace continuously until it is erased by temporarily lowering the potential of a " collector " electrode.

An important characteristic, more especially at high frequencies, is the signal-input impedence. One of the advantages of electric deflection, as with valve voltmeters, is that this impedance is so high that the effect on the signal circuit may often be neglected. It is in fact very similar to that of the negatively-biased grid of a valve : a capacitance of 4–10 pF in parallel with an almost infinite resistance.

Tubes for use where low plate capacitance is important have their plate connections brought straight out at the neck; otherwise, they are led down to the base. In practice all deflection plates must be tied to the final anode (other than p.d.a.) by resistances of not more than about 5 MΩ.

5.27. MULTIPLE-BEAM TUBES

The ordinary type of tube enables us to see how one quantity—say, voltage—varies with another—say, time. Sometimes it is helpful to be able to see how two or more different quantities vary with another; for example, the signal voltages across primary and secondary coils of an i.f. transformer with frequency, or current and voltage in a rectifier system with time. And for testing amplifiers with waves of various forms it is advisable to have the input signal as well as the output shown on the screen for comparison.

The simplest device is undoubtedly that employed in the single-gun double-beam tube originated by Cossor (Fig. 5.36) and now available in other makes—one extra plate midway between the Y plates and connected internally to the main anode and therefore at zero potential so far as the deflection system is concerned. At this stage the beam is undeflected and in a diffuse condition, so it is split in halves. When the split beam proceeds to come under the influence of the X plates, both halves are equally deflected by them. So if, for example, a time-base voltage is applied to the X plates, both beams are drawn horizontally across the screen; but they can be independently deflected vertically by the Y plates. If desired, one trace can be raised clear of the other on the screen by applying a suitable steady voltage in series with the signal voltage. Since the horizontal displacement is the same

* Hughes Aircraft Co., Los Angeles, Cal., U.S.A.

for both, any phase difference between the two waves is accurately depicted. It should be noted, however, that because deflection takes place each side of the plate, one of the pictures is inverted with respect to the other, and this must not be mistaken for a mysterious 180° phase shift!

This type of double-beam tube costs little, if any, more than the corresponding type of single-beam tube, and is interchangeable with it. There is a liability to " cross-talk " between the two beams, but this

Fig. 5.36—Deflection plates of double-beam cathode-ray tube. (A. C. Cossor, Ltd.)

is usually minimized by auxiliary electrodes. To give greater freedom from this complaint, and to allow independent X-deflection, etc., some tubes are produced with separate guns for the two beams. These advantages of course cost substantially more. But the method can be extended to more beams than two: an example by Twentieth Century Electronics has as many as eight. In the Electronic Tubes 4LP31 c.r.t. the split-beam type of two-trace provision is combined with post-deflection acceleration.

An alternative to a special tube, considered in Sec. 5.39, is to switch the Y plates to two or more signal sources in rapid succession.

5.28. RELATIONSHIP BETWEEN SIGNAL AND TRACE

When an alternating signal is applied between the plates of either pair the resulting trace is merely a straight line, which indicates the peak-to-peak *amplitude* of the signal, but not much else. If a signal of the same frequency and phase is applied also to the other pair of plates, the straight line becomes diagonal, the angle indicating the

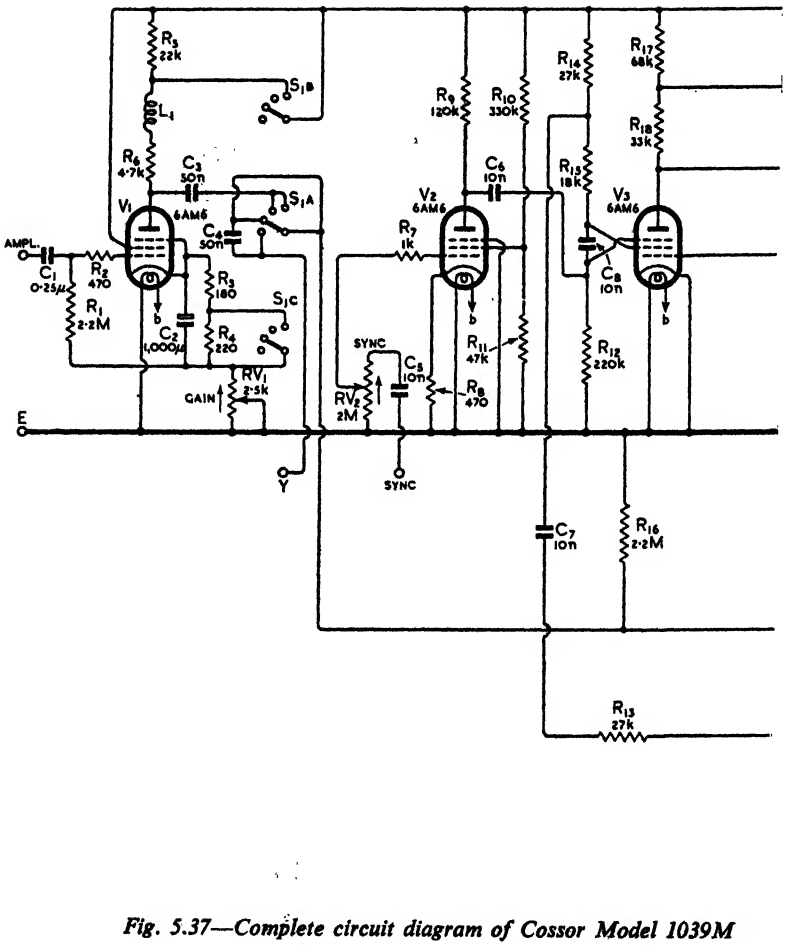
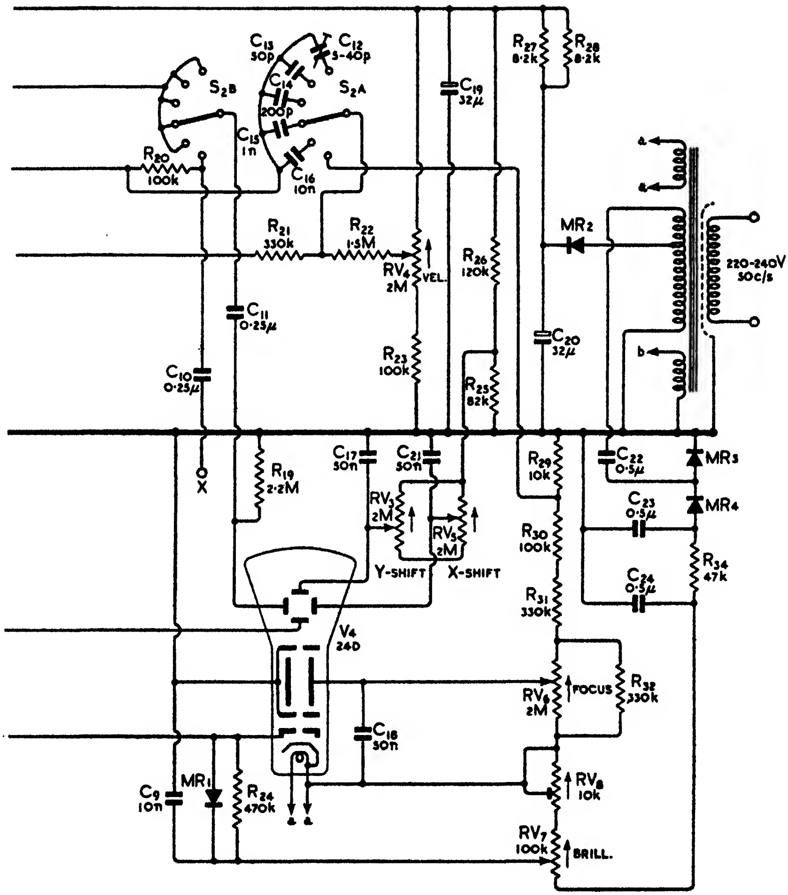


Fig. 5.37—Complete circuit diagram of Cossor Model 1039M



Mark II miniature oscilloscope. (*Cossor Instruments Ltd.*)

relative amplitudes; or if they are applied in different phase the line opens out into an ellipse or circle from which it is possible to deduce the *phase angle* (Sec. 9.10). By applying the input and output signals of an amplifier to X and Y plates respectively, it is possible to compare amplitude, phase, and also *waveform*; electrical distortion is shown by a visible distortion of the trace (Sec. 11.12). Again, two independent signals may be applied, and the *frequencies* can be compared with great precision (Sec. 10.10). In this case, one of the signals will generally be the "unknown" and the other from a laboratory oscillator or other calibrated source. In other problems it may be desired to study some signal, or quantity that can be converted into a signal, with respect to time; and it is for this purpose that the linear time base is designed. It is derived from an oscillator having a saw-tooth waveform, so that the spot is drawn across the screen at the desired rate, and then returned as rapidly as possible, to repeat the process with negligible loss of time. If the frequency of the time base is equal to the frequency of a continuously repeated signal wave or group of waves, the separate transient pictures of these always coincide and form a stationary figure which can be observed or photographed at leisure.

A most instructive book, with many beautiful photographs of actual traces, is *Cathode Ray Tube Traces*, by Hilary Moss (*Electronic Engineering*, 28 Essex Street, London, W.C.2).

Although all sorts of patterns or figures can be made to appear on the screen, it must be realized that they are optical illusions caused by the visual persistence of the eye, aided in some cases by the afterglow of the screen itself. At any one instant only a single spot of the screen is being touched by the ray.

5.29. POWER SUPPLIES

The one essential auxiliary is the power source. The heaters of c.r. tubes are similar to those in valves, but as it is usual for one of the anodes to be earthed the cathode is at a high negative potential, and as the potential of the heater must not be greatly different it is necessary for the heater current to be supplied from a separate highly-insulated winding on the power transformer.

Provision of the anode voltages is simplified by the trifling current demand. The anodes themselves take almost negligible current, usually a fraction of 1 mA, and the total loading is determined mainly by the chain of resistors from which the various electrode potentials are obtained. For most tubes the current drain need not be more than about 1 mA or so, and ordinary low-wattage components can be used. But of course insulation must be considered; such things as live grub-screws in control knobs must be covered up. Very simple smoothing is sufficient, even with half-wave rectification; and as the rectified voltage nearly equals the peak input an ordinary 350–0–350-V receiver transformer (so rated at a much heavier load) in series with a suitable half-wave rectifier gives nearly 1,000 V output—or, say, 900 V after resistance smoothing. In a simple oscilloscope described by the author (*W.W.*, Mar. 1950; reprinted as a leaflet), such a transformer is used to supply −900 V h.t. for the tube and also +700 V for the signal amplifier and time base. Alternatively, voltage doubling or multiplying rectifier circuits can be used, as in the Cossor miniature oscilloscope, Model 1039M Mk. II, the circuit of which is shown in Fig. 5.37. The potential divider is made up of R_{29}–R_{32} and RV_6–RV_8, with R_{34} for smoothing. Beam current, which determines brightness, is controlled by grid bias (RV_7), and focus is controlled by the potential of the second

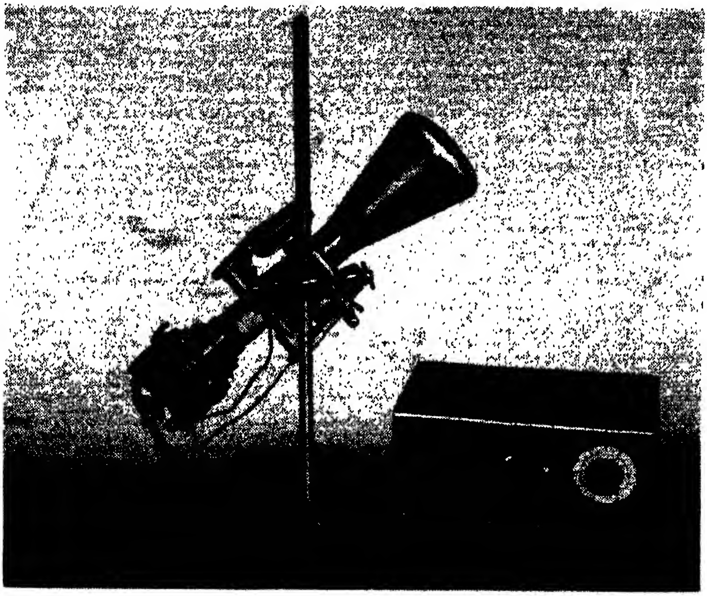


Fig. 5.38—A chemical retort stand is a convenient mounting for a cathode-ray tube, enabling the screen to be adjusted to a convenient angle for viewing, and the deflector-plate terminals to be brought close to the apparatus being tested. Note the 50-c/s deflector coils around the neck, with amplitude control at the base of the tube

anode, RV_6. In most tubes its voltage is somewhere about one-fifth that of the final anode. The tube shown has the first anode joined internally to the third, but this is not so with all types.

The final anode, other than p.d.a., is usually earthed, and the potentials of the deflection plates are reckoned relative to it. All the plates must have conducting paths between them and this anode, such as R_{16} and R_{19} in Fig. 5.37, otherwise the trace is liable to stray off the screen. A great convenience is to be able to move the spot up and down, and sideways, independently of the deflecting signals; these facilities are called Y shift and X shift respectively, and are provided by RV_3 and RV_5. To enable the shifts to work both ways from centre, the negative potential of the ends joined to R_{30} is matched by a positive potential from across R_{25}. Double-beam tubes need separate Y shifts for the Y plates. In the more refined oscilloscopes the shifts are directly calibrated in volts and seconds, enabling the voltage and duration of a signal or of any part of it to be read off by shifting the trace past a fixed mark. (See Secs. 5.32 and 5.36.)

In laboratories a c.r. tube is so rarely seen outside an oscilloscope comprising (at least) a power unit, time-base generator and signal amplifiers, that one can fail to realize that a complete oscilloscope is not always essential nor even the best setting for c.r. oscilloscopy. For many purposes its auxiliary circuitry is not only unnecessary but a positive disadvantage, adding stray coupling and self-capacitance to the deflection plates. In such cases a convenient support for the tube is a chemical retort stand (Fig. 5.38). This enables it to be set at the most convenient angle and its deflection plates connected to the test points more directly than is possible with the conventional oscilloscope. The power unit, being connected by a flexible cable, can be kept several feet away from the tube, so that the well-focused spot is not drawn into a line by stray magnetic field from the power transformer. If a transformer must be placed close to the tube, it must be protected by a mumetal shield, and other precautions taken. This by no means simple problem is discussed by W. Tusting (*W.W.*, Dec. 1951).

5.30. DEFLECTION AMPLIFIERS

Most c.r. tubes, even with the lowest satisfactory anode voltage, need something of the order of 25 V (peak to peak) between the Y plates for 1 cm deflection. Exceptional tubes have 10 or more times greater sensitivity; for example, the Ferranti 5/63 series (Sec. 5.26) and the Electronic Tubes 5CLP31. This is most commendable, because amplification is a thing to be avoided if at all possible. The reason is not so much the extra apparatus needed as the risk of introducing undesired effects—variation of gain and phase-shift with frequency, and non-linearity, for example. To maintain the gain down to zero frequency invites drift, as we saw in Sec. 5.19. To extend it up to very high frequencies the shunting effects of stray capacitances must be evaded by using low-resistance couplings, and doing that restricts the output voltage that can be obtained with satisfactory linearity, unless an inordinate amount of power is consumed. Modern wide-band oscilloscopes usually have to incorporate an electric fan to carry away the heat. And it is futile to use a c.r.o. to study distortion if the amplifier is putting in distortion of its own.

For a restricted frequency range—say the a.f. band and perhaps low r.f.—these difficulties are less acute, so a common device is to provide high gain for this range and a lower gain effective up to much higher frequencies and perhaps down to zero. Using conventional video technique, the top frequency can be pushed up to about 40 Mc/s. Above that, distributed amplification—in which a number of valves are operated in parallel but separated by sections of transmission line so that their shunt capacitances do not add up—is used to extend the top limit to about 100 Mc/s. Using special wide-band valves, such as the Mullard E810F with a g_m of 50 mA/V, similar performance can be obtained with conventional amplifier circuits. A gain of 2 or more per stage is possible with load resistance of the order of 50–70 Ω; e.g., directly across coaxial cables.

The frequency limits in amplifiers are usually defined as the frequencies at which the response has fallen 3 dB from the main "plat-

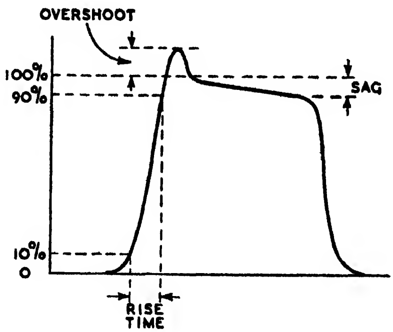


Fig. 5.39—Amplitude/time graph of a pulse, as seen on an oscilloscope, defining the quantitative meanings of standard terms

eau"; that is to say, by nearly 30% in voltage. Since the object of having a wide frequency band in oscilloscopes is often to preserve pulse waveforms from distortion, the information is more helpfully given in the form of "rise time", which is the time taken for the spot to be deflected from 10% to 90% of its amplitude when the pulse fed in has an instantaneous rise (Fig. 5.39). Thus in the Tektronix Types 581 and 585 oscilloscopes, which employ distributed amplifiers, the speed of Y deflection is stated alternatively as 100 Mc/s and 3·5 mμ sec (or n sec) rise time.

In the effort to increase deflection speed another difficulty arises. If the electron beam spends an appreciable fraction of a signal cycle between the Y plates the signal voltage will have varied appreciably during that interval, and may even have begun to reverse. The treatment is to speed up the electrons by higher anode voltage, and to reduce the length of the Y plates, but both these measures reduce the deflection sensitivity and necessitate more amplification.

If identical waveforms are repeated at short intervals, the high-speed problem can be dodged and the equivalent of a bandwidth of several hundred Mc/s obtained by the sampling technique. An example (in the four-figure price bracket) is the Hewlett-Packard Model 185A/187B: 0 to 1,000 Mc/s. The waveform seen on the screen actually consists of perhaps 100 fragments extracted from that number of successive signal waveforms. The principle is basically the same as that used in the stroboscope for viewing high-speed machinery in slow-motion.

If the signal pulses themselves are used to trigger the time-base sweeps, and are delayed by short but precisely equal times before being applied to the Y plates, there is no need even for the intervals between pulses to be equal. The Mullard Type L362 oscilloscope, which is operated by transistors, works on this principle.*

The same frequency range (0 to 1,000 Mc/s) can now be covered directly, without sampling, so that it is available for " one-shot " waveforms. This is accomplished by the technique of distributed deflection, in the Tektronic Type 519 oscilloscope. The Y plates are divided into

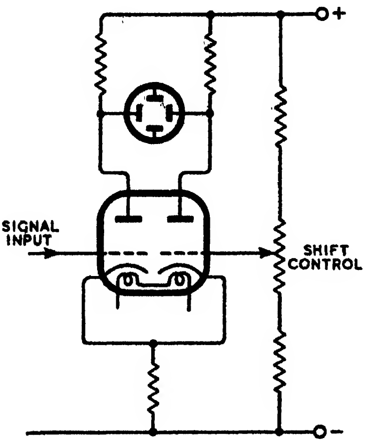


Fig. 5.40—A commonly-used type of circuit for applying balanced push-pull signals to deflection plates

a number of sections in the direction of the beam, linked by small inductors. The signal is introduced at the end nearest the cathode and travels to the other end—which is terminated by resistance to prevent reflection—at the same speed as the electrons in the beam, so that deflection is in phase and cumulative. The same idea is the basis of the travelling-wave c.r. tubes by 20th Century Electronics and G.E.C.

Deflection amplifier circuitry is too varied to describe here in detail; but direct coupling and anti-drift precautions are obviously needed for z.f. amplification, and low shunt capacitances and coupling resistances and high mutual conductances for high frequencies. Design information on how to achieve a top frequency of 30 Mc/s is given by G. H. Leonard (*W.W.*, Aug. 1958). If, in order to avoid the z.f. drift problem, stages are connected through series capacitors, their capacitances have to be large in order to avoid phase shift and sag (Fig. 5.39) at low frequencies, and then there are irritatingly long waits for the trace to take up its position while they charge or discharge following a change in the d.v. component, to say nothing of the inability to measure such components. For the same reason, shift controls also should preferably be arranged so that blocking capacitors (as in Fig. 5.37) are not needed. Push-pull output is used in all but the

* Described more fully by G. B. B. Chaplin, A. R. Owens and A. J. Cole in a paper in *Proc. I.E.E.*, Vol. 106, Part B, Supp. 16, May 1959.

cheapest oscilloscopes, for the sake of its linearity and doubled amplitude and because most tubes benefit by symmetrical deflection. Both features are combined in the popular arrangement shown in Fig. 5.40. If direct coupling is used throughout, for frequencies down to zero—and that is most desirable—push-pull is usually used throughout, to minimize drift.

It must be remembered that deflection is caused by a difference of potential *between* a pair of plates, not by their potential relative to anything else. But even in tubes designed for asymmetric deflection it is a good thing to have as little average p.d. as possible between either pair of plates as a whole and the final pre-deflection anode, otherwise the plates act as unauthorized anodes, distorting the field pattern and causing the form of defocusing known as astigmatism.

Gain control requires more thought than one might expect. It is needed not only to fit signals of widely different amplitudes to the size of the screen but (in some oscilloscopes) to adjust the calibration of fixed voltage scales. Adjustment ought not to introduce or alter a d.v. component, even in capacitor coupled systems, where transient disturbances of the trace may last too long for one's patience! One of a number of devices for continuous gain control is to substitute for the single cathode resistor in Fig. 5.40 separate double-resistance ones for each valve, and have a gain-control variable resistor between cathodes. Stepped controls usually consist of potential dividers, as discussed in the following sections. In general it is undesirable for the gain control to affect the bandwidth and hence the waveform seen on the screen. In some circumstances, however, one may be glad to trade gain for bandwidth. For example, control of negative feedback does just that. And in some oscilloscopes there is one wide-band low-gain amplifier (perhaps direct-coupled) always in circuit, preceded by one of narrower bandwidth when high gain is necessary.

Because such varied needs have to be met by an oscilloscope, and good ones are expensive enough without one's having to acquire a whole battery of them for different purposes, some are made adaptable by plug-in units, especially Y amplifiers.

In the simplest models no X amplifier is provided, but one or both X plates are switchable to terminals for external access. An X amplifier is very desirable, however, not only for such externally connected signals but also for providing symmetrical deflection, for facilitating X shift, and for improving time-base linearity as described in Sec. 5.33.

5.31. INPUT ARRANGEMENTS

Provision is usually made for switching the amplifier out when the signal is strong enough to give adequate deflection direct. To cope with signals that are too strong, an attenuator is needed. In some models the amplifier is in circuit all the time, and by means of a switched attenuator the overall gain is varied in steps.

Attenuators consist basically of high-resistance potential dividers. At high frequencies, stray capacitance shunts the resistances by reactance, so to ensure that the attentuation is not affected it is necessary to supplement these strays by fixed or pre-set capacitors to ensure that the time constant ($= RC$) of every resistor is the same, including the input of the amplifier (or c.r.t.). That is usually a few MΩ shunted by a few pF. Any comparatively low resistances in the attenuator have to be shunted by proportionately higher capacitance. Fortunately these are not the ones that would appear directly across the input, but by

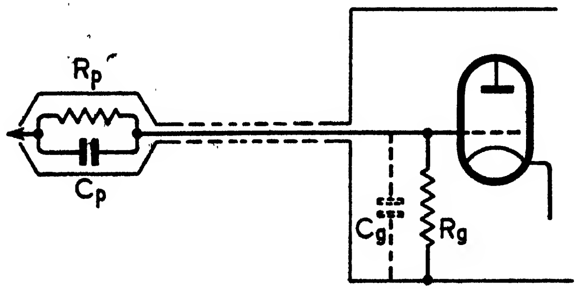


Fig. 5.41—Attenuating probe, which gives the same attenuation at all frequencies if C_p is chosen so that $C_pR = C_gR_g$, C_g being stray capacitance

the time the capacitance of wiring, attenuator switch and connecting cable have been added in, the input admittance amounts to a formidable load at high frequencies. To reduce it, a probe is often supplied. One form consists simply of a high resistance (R_p in Fig. 5.41) shunted by just enough capacitance (C_p) to equalize the time constant to R_gC_g. R_p would be chosen to give a known attenuation, say 10 : 1, in which case the input capacitance would be reduced to one tenth. This is often worth while, even if it means having to increase the amplification. An alternative is to use a cathode follower in the probe; it causes hardly any attenuation, but if an extra valve is being used in the interests of low input capacitance it is more convenient to put it in the body of the oscilloscope, as an amplifier to compensate for the loss in a simple attenuator probe. However, a cathode-follower probe, using an electrometer valve in a special stabilizing circuit to give a phenomenally high input resistance (current input 2×10^{-13} A) is described by G. O. Crowther (*E.E.*, June 1955). It is, of course, equally applicable to a valve voltmeter.

Although voltage is the usual signal parameter observed by c.r.o., current waveforms can also be studied by precisely the same methods as with meters. In particular, clip-around probes are obtainable for observing current waveforms without breaking the circuit (Sec. 5.11).

5.32. VOLTAGE MEASUREMENT

The tendency is for at least the better-class oscilloscopes to provide for the signal voltage to be indicated. One policy is to regard the c.r.o. as a variety of electronic meter and provide the screen with a voltage scale (graticule) and the Y amplifier with an attenuator switch, varying its gain in steps of (say) 1–3–10, etc. The shift control enables any point on the waveform to be aligned with zero on the scale. This system is most convenient in use, but also liable to considerable error, because it assumes that the gain of the amplifier on each range is always the same, at all frequencies, and that the deflection sensitivity of the c.r.t. is always the same (which necessitates anode voltage stabilization) and that amplifier and tube together have a linear deflection/input-voltage law. Even with appropriate precautions no better than 10% error is claimed for one example of this type (Solartron CD.523.5).

The alternative is to apply some kind of calibrating voltage to the input of the amplifier; this allows for any gain variations other than those rapid enough to occur between calibration and reading. Ideally, the calibrating voltage would always have to be of the same frequency as the signal, but that is rather a lot to expect of a built-in calibrator! Provided that the gain of the amplifier is reasonably level down to z.f.,

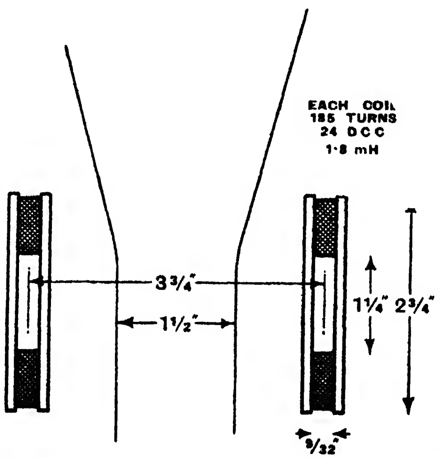


Fig. 5.42—Dimensions of coils suitable for providing a uniform 50-c/s deflecting field

the most suitable calibrating " signal " is a known shift voltage applied to the input. The part of the waveform to be measured is compared with the amount of shift voltage giving an equal vertical movement on the graticule. A popular method is that seen in Fig. 5.40. (There may of course be other stages between input and c.r.t.).

5.33. TIME BASES

For examining the variation of a voltage with respect to time, a voltage representing time is applied to the X plates. For most purposes it is convenient if this voltage varies linearly with time; i.e., is directly proportional to it. But occasionally it may be sufficient to use the much more easily provided sinusoidal voltage, derived from the 50-c/s mains if that frequency gives a suitable rate of change. This is the purpose of the pair of coils seen in Fig. 5.38, energized from a heater winding on the power transformer and controlled by a potentiometer of, say, 10 Ω. Dimensions of suitable coils are given in Fig. 5.42, but there is no need to keep strictly to these.

This deflector is used at the start, during focusing and brilliance adjustment, because a stationary spot is very bad for the screen and does not give a fair idea of visibility. Then it constitutes a 50-c/s frequency standard, when the mains are time controlled, and can be used for calibrating frequency. And it serves as a time base whenever a strictly linear law is not essential. It can sometimes be used even when a linear deflection is wanted, because the " middle cut " of an extended sine-wave trace is quite a reasonably good approximation to it. It is necessary merely to increase the size of the base line so that the ends are well off the screen (if the coils are not powerful enough it can be done by applying perhaps 350 V a.v. to the X plates) and then to arrange that the Y-plate signal, which is to be observed, occurs near the centre. In this manner, waves of the order of 1,000 c/s can be examined on a 50-c/s base.

More generally useful is a sawtooth time-base waveform, in which the voltage increases linearly at a desired rate, deflecting the spot horizontally across the screen, and then returns as rapidly as possible to the starting point. During the return (" flyback ") the beam should be cut off by negative grid bias. For many purposes these sawtooth cycles are required to repeat continuously, so that successive signal cycles (or groups of cycles) are superimposed on the screen to form a single steady trace. For examining non-repetitive or random forms, however, time base cycles must be triggered only when required. A vast variety of linear time-base circuits have been devised, and in fact a book about the same size as this is devoted wholly to them—O. S. Puckle's *Time Bases*, 2nd edn. (Chapman & Hall, 1951).

The basic principle of nearly all of them is to charge a capacitor through some constant-current device, and then when it reaches a certain voltage to discharge it, the faster the better. By controlling the capacitance or the charging current, or both, the velocity of the spot across the screen can be varied as desired. If the flyback time is made negligible and the next sweep starts without delay, the velocity of the spot, or the time of its working sweep, is simply related to the frequency of the time-base generator; but, as these conditions do not always apply, the tendency in specifications is to state the available ranges of time per cm or per screen width.

Thyratrons were once much used for time bases, because they pass negligible current until a certain voltage (easily adjustable) is reached, when they suddenly conduct very freely indeed, discharging the capacitor quickly and so providing a rapid flyback. But the time of recovery from this condition begins to be a serious fraction of the cycle

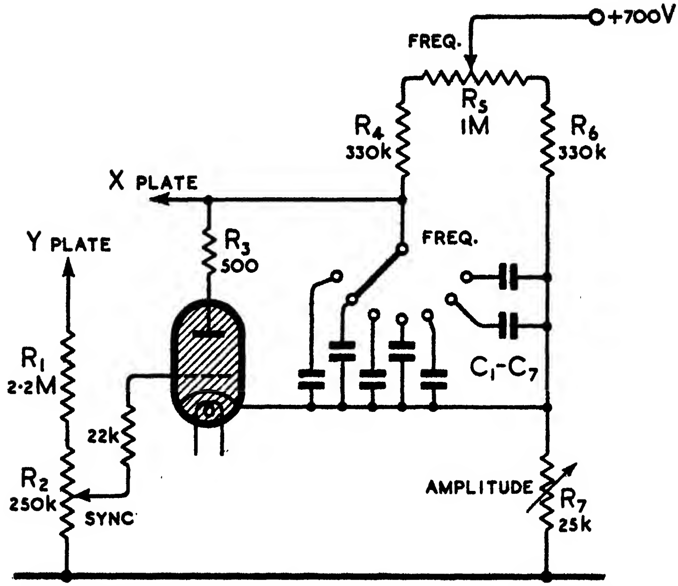


Fig. 5.43—Simple practical time-base circuit for a.f. and low r.f.

at a few tens of kc/s and restricts the top frequency. In length of life and constancy of characteristics, thyratrons are less good than hard valves. So they have gone out of use in commercial oscilloscopes. But if one wishes to make a simple time base for a.f. and low r.f. the thyratron is the best single-valve choice of all for ease of control and speed of flyback. Fig. 5.43 shows a practical circuit. The ranges of frequency given by C_1 to C_7 are approximately as in Table 5.2. The

Table 5.2

Capacitance, nF		Frequency, c/s
C_1	300 (= 0·3 μF)	12–53
C_2	100	35–160
C_3	30	120–530
C_4	10	350–1,600
C_5	3	1,200–5,000
C_6	1	3,600–14,000
C_7	0·3	13,000–40,000

rate of charging is approximately constant even though an ordinary resistance (R_4) is used instead of a pentode valve or other constant-current device, because only about one sixth of the available charging voltage is actually needed for deflection in a typical c.r.t. with moderate anode voltage. R_6 is for passing a compensating current through R_7 (controlling the bias) so as to minimize variations in amplitude when the frequency is controlled by R_5. R_2 is to enable a small controllable fraction of the " work"—the Y signal— to be impressed on the grid of the thyratron for synchronizing the time sweeps so as to keep an exact whole number of work cycles on the screen.

This (and other time-base circuits) can be improved by push-pull output, as in Fig. 5.40. The amplification enables a still smaller proportion of the charging voltage to be used, giving better linearity, even with a lower voltage than in Fig. 5.43.

Too many different hard-valve time-base circuits are now in common use to describe fully here, and compared with thyratron circuits they are more complicated—some of them much more complicated—because special devices are required to initiate the flyback discharge and to force it through at high speed.

The Puckle circuit was the first and for some years almost the only one in use, but has been generally superseded by others capable of greater speed, linearity, etc. It charges the capacitor through a pentode (as a controllable high resistance), and discharges it through a triode coupled to a third valve to obtain a trigger effect. The cathode of the triode is unfortunately at sawtooth potential.

The basis for many modern time-basis circuits is a linearizing circuit usually called the Miller integrator—inappropriately since it was not invented by Miller but by A. D. Blumlein, who also developed it into a time-base generator. Fig. 5.44 shows the essentials. R is large

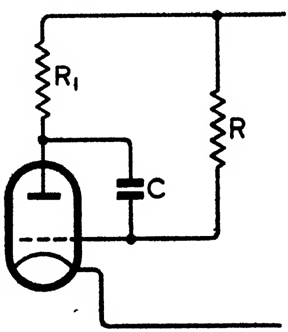


Fig. 5.44—Miller—or Blumlein—integrator circuit, in which a closely constant rate of charging C is obtained. This is the basis for much-used time-base and time-delay circuits

compared with R_1. Assume that C is charged to the full h.t. voltage. It will then begin to discharge through R_1 and R, and owing to the drop across R_1 the grid will be negative. As anode current starts to flow it increases the drop across R_1, making up for the falling voltage across C. The voltage across R is thereby kept very nearly constant, which means that the current through it and hence the rate of charge is constant. Directly grid current flows, however, the action ceases. With a suitable pentode a large anode voltage sweep is obtained for a very small grid voltage change.

What is needed to develop this device into a time base is some means for periodically recharging C. The simplest method is to make the same valve perform as a transitron oscillator, as in Fig. 5.37, where V_8 is the valve concerned. Most of the time the suppressor grid g_3 is near zero potential, but as C (one of C_{12}–C_{16}) nears the end of its discharge, the pentode " bottoms " and current is diverted from anode to g_2, causing a sharp fall in its potential. This is passed on to g_3 by C_8, making it negative enough to cut off anode current and allowing C to recharge through $R_{17} + R_{18}$ and the grid-to-cathode path. Meanwhile C_8 is charging and g_3 is becoming less negative, allowing anode current to flow once more and reduce g_2 current. This accelerates the positive movement of g_3 potential, and the initial state is restored. Note that the X sweep is taken from across R_{17}, and beam suppression during flyback is obtained by applying the pulse across R_4 to the c.r.t. grid.

With a fixed value of g_2–g_3 capacitance, as in Fig. 5.37, the flyback time is more or less constant, which means that it is unduly long for the higher-frequency sweeps. So its value is often changed along with that of the main C by another wafer on the frequency switch. Even then it is not a really fast flyback. For improved performance in that and other respects, a variety of circuits have been evolved by the addition of one or more valves and diodes. Similar circuits are also sometimes used for obtaining a precise and adjustable time delay. A circuit combining the exceptional linearity of the Blumlein integrator with the flyback system of the Puckle time base has been described by C. S. Speight (*W.W.*, Jan. 1959).

Another basic linear circuit is that known as the bootstrap. Unlike Blumlein's, in which the load is in the anode circuit, it is a cathode follower, and C is joined to earth instead of anode. The rising grid voltage, as C charges, pulls up the cathode potential and offsets the rise across C so far as the charging voltage is concerned. In the Marconi Instruments TF.1330, which is a high-performance oscilloscope, a combination of Blumlein and bootstrap circuits is used. Transistor versions of both are discussed (though mainly with a view to television) by K. P. P. Nambiar (*E.E.*, Feb. 1958).

Other more or less linear time-base circuits, found mostly in the simpler oscilloscopes, are multivibrator or blocking oscillator types, similar to those in television receivers except that they are designed for voltage instead of current deflection.

A circuit which, although using three valves, is particularly suited to the amateur constructor and gives a good general performance, is the " grid-diode " type described by T. A. Mendes (*W.W.*, Dec. 1957).

5.34. SYNCHRONIZATION

In its commonest role an oscilloscope displays signal waveforms repetitively on a linear time base as an apparently stationary trace. It is not practicable to adjust the frequency of a free-running sawtooth generator to achieve this condition and maintain it without some form of coupling between the signal and the generator to synchronize the latter. Simply connecting an active Y plate through a high resistance and/or small capacitance to the grid or other sensitive point in the time-base circuit usually succeeds after a fashion, and has been much used in the simpler oscilloscopes. But there are certain disadvantages. Even a small amount of coupling may load the signal source enough to distort the waveform appreciably. The coupling has to be adjusted whenever signal amplitude or frequency are changed, or the trace will go out of synchronism or close up at one end. And there is no choice of the part of waveform at which triggering takes place.

In the interests of convenience and versatility the synchronizing system is now often quite elaborate. Even the simple Fig. 5.37 circuit has one of its three valves (V_2) allocated wholly to this duty, in which it acts as a buffer amplifier. To improve the precision of triggering—very necessary, when time is measured in millimicroseconds, to prevent jitter—a pulse-shaping stage is sometimes included. Choice of positive or negative going parts of the waveform may be provided, for example, by switching to either Y plate with symmetrical deflection. Choice can be further extended by means of a variable trigger bias. Such facilities are particularly valuable in examining television and other pulse waveforms. Where signal cycles follow one another without breaks, it is usual to synchronize to a submultiple of the signal frequency to ensure that the whole of at least one cycle is fully displayed in any phase.

The foregoing descriptions assume that the time base would continue running even if there were no signal—as in a television receiver. To enable single waveforms to be displayed, provision is sometimes made for the sawtooth generator to be put into a " one-shot " condition, in which sweeps take place only when triggered. This state could of course be used for examining repetitive waveforms too, but would conceal any signals too small in amplitude to trigger the time base. To enable the whole of a transient waveform to be seen, it is necessary to delay it between the point at which it does the triggering and the Y plate. An artificial delay line is used for the purpose, and 0·25 μsec is a typical delay, allowing a high-speed time base to be under way by the time deflection takes place. The trace is preserved for examination by a long-afterglow screen phosphor or by photography (Sec. 5.41).

5.35. TRACE EXPANSION

Even in such everyday work as television receiver servicing there is need to examine parts of the signal waveform too small to show up clearly when the whole is displayed. The most straightforward solution is to increase the amplitude of the time base beyond the width of the screen if necessary, using X shift to bring the desired portion into view. This is of course only possible if there is a substantial excess of amplitude available, and is one more reason for having an X amplifier. The Fig. 5.40 circuit is suitable. An intermediate stage may be required if the time-base output is small. Fortunately there is no need for linearity to be preserved beyond the bounds of the screen; the circuit just mentioned in fact cuts off fairly sharply beyond a certain input voltage. Ample X shift voltage must be provided.

That method is workable up to perhaps a fivefold expansion and is simple and effective. Where much greater expansion is needed—e.g., picking out a single line from a television raster—the usual means is a precise and adjustable time delay circuit, commonly based on the Blumlein integrator. The sequence of operations is that this circuit is triggered by a suitable feature of the signal waveform and starts to " run down "; after an interval sufficient for the signal waveform to approach the desired section is reached a level, set by the delay control, at which it triggers a one-shot time base, which has been set to an appropriately high speed to cover only the desired section, at the end of which the beam flies back to the start and awaits the next triggering pulse. Obviously the whole of this circuitry must be highly stable and precise if the trace is to be steady.

5.36. TIME MEASUREMENT

At any one setting of a truly linear time-base generator the distance travelled horizontally by the spot is directly proportional to time, so the screen could be fitted with a linear time scale, enabling the duration of a whole cycle (and hence the frequency) or any part thereof to be measured. Suitable arrangements for changing the scales to fit the signal to the size of the screen and for ensuring their accuracy might seem to be the same as those for vertical deflection (Sec. 5.32), but there is an important difference in the problem, apart from the usually wider range to be covered. Whereas the precise vertical amplitude is unimportant, the horizontal amplitude occupied by a cycle of a repetitive signal must be a submultiple of the time-base amplitude. So if fixed time scales are used the time-base velocity must be varied only in fixed steps to fit them, and synchronism obtained solely by varying time-base amplitude, which is rather untidy unless a large number of velocity steps are provided. Modern time-base circuits can be made sufficiently stable and linear for time calibration, but there must of course be the usual precautions to keep the deflection sensitivity constant, including any amplification that may be used.

An alternative—or additional—scheme is to provide a time calibrating signal from what is essentially a standard-frequency generator. An ordinary external frequency-calibrated sine-wave oscillator can be used at a pinch, the distance between each pair of successive peaks on the screen representing a time equal to one period of oscillation. Time comparison is more precise and convenient if pulse-shaping circuits are used to charge the waveform into something like Fig. 5.45. An alternative presentation is by means of brightness modulation of the trace, produced by applying these " pips " to the c.r.t. grid.

If the calibration oscillator ran continuously, its markers would not, in general, synchronize with the signal. It is therefore necessary to stop and restart them for each time-base cycle. This practically rules out the obvious choice of a crystal oscillator as a reliable time calibrator. A commonly chosen one is the " ringing " oscillator, which comprises a high-Q stable oscillatory circuit and means for shock-exciting it by (or coincident with) the bright-up pulse that makes the spot visible

Fig. 5.45—For a time-calibrating scale, pulses of some such form as this are more convenient and precise than plain sine waves

during the sweep. Fig. 5.46 shows a common arrangement in which the LC circuit is excited by the valve current being suddenly cut off at the start of the sweep. At the end, current is restored and the low cathode impedance of the valve quickly damps out the tail of the oscillations.

5.37. THE POLAR TIME BASE

Although the straight horizontal time base so far described, giving screen patterns based on the cartesian co-ordinates of mathematics, is

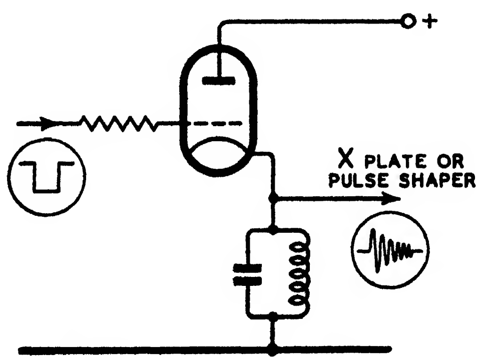


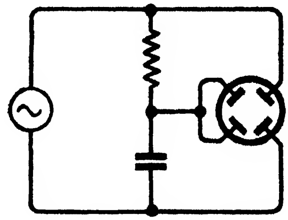
Fig. 5.46—Simple form of ringing oscillator for producing synchronized trains of oscillations

much the commonest, it is limited in extent by the diameter of the screen. By adopting polar co-ordinates, the base takes the form of a circle and is continuous. It is generated by applying to both X and Y plates alternating voltages of the same frequency but in phase quadrature (90° different), which can be done very simply as in Fig. 5.47. R and C should be such as to give equal X and Y deflections. To make the signal produce radial deflection it is applied in series with the anode. An example of this technique is given in Sec. 10.10.

A partial use of this method is sometimes helpful in overcoming one disadvantage of the ordinary sinusoidal time base—that the go and return traces are superimposed, so that it is difficult to count the number of cycles. If a relatively small quadrature component of the X-plate voltage is applied to the Y plates along with the signal, the traces are more easily distinguished, as in Fig. 5.48.

A special tube, described by von Ardenne (*W.E.*, Jan. 1937), enabled a spiral time base 160 in. long to be displayed on a 4-in. screen! Similar

Fig. 5.47 — Phase-splitting circuit for generating a circular time base



results can be contrived with ordinary tubes by modulating the a.v. shown in Fig. 5.47 with a saw-tooth waveform.

There are problems in which the base should be some quantity other than time. There is no end to what can be done by translating various quantities into deflecting voltages, and so reproducing performance curves on the screen. Valve characteristic curves, for

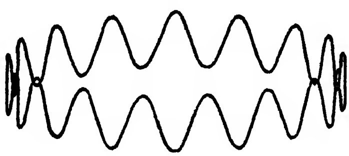


Fig. 5.48—Signal display on sinusoidal time base, "opened out" by addition to signal voltage of 90° displaced time-base voltage

instance, can be shown instantly on the screen instead of tracing them out laboriously by meters. Not only is there the saving in time but it is possible to investigate portions of the characteristics at which the valve could not safely be held long enough to take a meter reading. Such apparatus is described in Sec. 9.37.

5.38. THE FREQUENCY BASE

One non-time base outstanding in importance is the frequency base, used for observing a.f. or r.f. frequency characteristics of transformers, amplifiers, tuners, filters, or complete receivers, as an alternative to the laborious process of plotting them point by point with a variable-frequency signal generator and a valve voltmeter. What is needed is a proportionate relationship between X deflection and the frequency of the signal generator.

In practice X deflection is assumed proportional to time-base voltage, so a usual approach is to control the frequency of the generator by the time-base output, as shown in Fig. 5.49, where the scheme is applied to the overall r.f. selectivity of a receiver. Several devices for making frequency proportional to voltage are mentioned in Sec. 4.36. It may be noted that time-base linearity is quite unimportant; in fact, in equipment made for this specific purpose, plain 50 c/s from the mains has sometimes been used instead of a time-base generator, and the deflecting agent in Fig. 5.49 is only labelled a *time* base because it is convenient to use if it is available, and gives the most uniform brightness of trace. The same 50 c/s supply can be used to vary the frequency by a mechanical device, or a mechanically rotated linear tuning

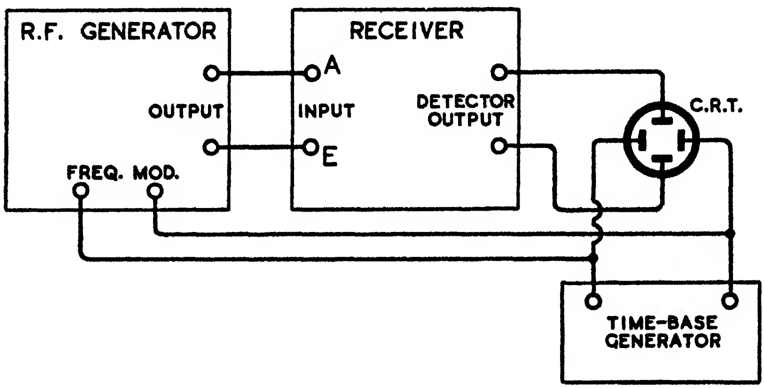


Fig. 5.49—Block diagram of apparatus for frequency-modulating a r.f. signal generator by the time-base voltage, for displaying a continuous selectivity curve of the receiver on the c.r. tube

capacitor can be arranged to trigger the usual linear time base once per revolution by a contact on the shaft. Mechanical methods are sometimes useful as a temporary lash-up, but generally electrical methods of " wobbulation " are preferred.

Whatever the method, care must be taken that the wobbulation is not so rapid that the apparatus under test is unable to respond quickly enough. Using an ordinary fast c.r.t. screen, a frequency of about 10 sweeps per second is needed for the eye to take in the response; with a.f. tests, and r.f. sets containing very sharply tuned circuits, a sweep time of 1 to 10 seconds is needed, using a long-afterglow screen or a camera.

5.39. MULTIPLE TRACE DISPLAYS

The ability to display two traces simultaneously for comparison is very often most helpful, and for some purposes a larger number is needed. One solution is a modification of the tube itself; split-beam and multiple-beam tubes have already been considered in Sec. 5.27. Another is to use several separate tubes, which would normally be run from a common power supply and perhaps a common time base. If many tubes are used, small ones are naturally preferred, sometimes with 1-in. screens. In the compact Airmec Type 249 oscilloscope there are four, each a little over 2 in. in diameter. The Mullard D–16–22 tube,

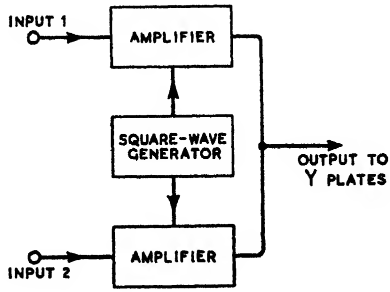


Fig. 5.50—Principle of beam-switching device for enabling two traces to be displayed on an ordinary single-beam tube

having a screen 5½in. × 1½in., is particularly suitable for grouping.

The other main alternative is to use a single ordinary tube with a single beam and share it in quick succession among as many signals as are required to be displayed. Some oscilloscopes with single-beam tubes are (or can be) provided with electronic switches for performing this service. The general scheme for a double trace is shown in Fig. 5.50. Although this looks simple, it has to be carefully designed if it is to give satisfaction. The square-wave generator is usually a multivibrator or similar circuit. Its outputs are applied to the suppressor grids of the amplifiers, or in some other way made to cut off one amplifier with the negative half-cycle while the positive half-cycle is holding the other amplifier in the active state. Clearly it is necessary to ensure that the gain remains constant during this period, and that no stray coupling prevents cut-off of the other being complete. Provision must be made either for switching Y shifts too, or for introducing a trace-separating voltage between the two outputs.

The switching frequency needs to be considered. If the time-base frequency is high, one may decide to allow a whole sweep—or even several—belonging to one channel to occur during each switching half-cycle. Synchronizing the switch with the time base is not essential, but is likely to give a better result, because switching occurs during fly-back. On the other hand, for low t.b. frequencies, with which the previous system would accentuate flicker, the switching frequency can be comparatively high, breaking up the traces into dashes (or dots, if very high), but if the frequency is random these may not show. Some design information has been given by K. R. Sturley (*W.W.*, Jan. 1953).

Using a 10-cathode neon counter tube, K. E. Wood and T. C. Keenan have extended the beam switching principle to give nine traces

(*E.E.*, Mar. 1956) but the maximum input frequency is only 100 c/s. As many as 24 traces have been presented—with the ability to expand any section of the sweep—on a 21-in. television tube (*E.E.*, Jan. 1956, p. 38), up to 500 c/s on each.

An entirely different Y-deflection technique which is capable of displaying two or more traces simultaneously has been shown by R. J. D. Reeves (*W.W.*, Feb. 1956; "Voltage Coincidence Oscilloscope"). Y-deflection is not by the signals at all, but by a continuously running oscillator which sweeps the beam rapidly up and down across the screen.

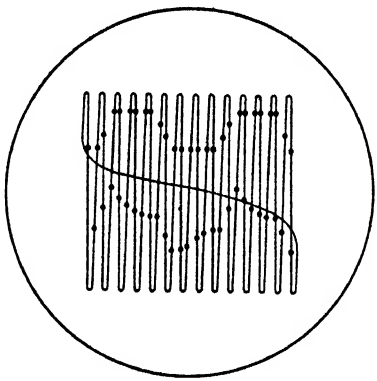


Fig. 5.51—Voltage-coincidence system of displaying more than one trace at a time on a single-beam tube. The continuous raster is normally made invisible by adjustment of c.r.t. bias so that only the dots outlining the traces can be seen. The gaps between the dots are filled in by rapid succession of unsynchronized rasters

If it were visible it would look like the continuous sine waveform in Fig. 5.51, but actually it is blacked out except where its voltage coincides with that of any of the signals. Such coincidences are made to reduce the c.r.t. grid bias and display the signals as a series of dots. Since the X and Y frequencies are unrelated, the dots due to successive sweeps tend to fill in the gaps and (if the sweep frequency is not too low and the afterglow too short) build up a continuous trace of each signal waveform.

5.40. COMPLETE OSCILLOSCOPES

Some idea will have already been gathered of the wideness of the price and facility range of commercial oscilloscopes. The Cossor Model 1039M Mk. II, whose circuit diagram is Fig. 5.37, is a miniature oscilloscope with only three valves, intended primarily for servicing domestic receivers. A design by the author (*W.W.*, Mar. 1950; reprinted as a leaflet with the wobbulator mentioned in Sec. 4.36) has only two valves, yet is capable of much useful work up to a few hundred kc/s. At the other extreme are models with nearly 100 valves and four-figure prices. The era of these highly sophisticated instruments may be said to have begun with the first Tektronix oscilloscope, which set a new standard of workmanship and performance. Wide frequency range of Y amplification, direct reading of voltage and time, and precision and versatility of control, are among the features of this class of instrument. The Tektronix layout, with the tube screen in the top left-hand corner, tube controls on the right, and plug-in amplifiers below the tube, has become almost standard for expensive oscilloscopes. For details of instruments at both these extremes and all between, the makers' literature should be consulted, and also a valuable review, " Modern Oscillosope Practice " (*E. & R.E.*, June 1958). A most interesting review of the considerations behind the design of a versatile general-purpose oscilloscope (Marconi Instruments TF1330) appears in *M.I.*, June 1959, supplemented in Dec. 1959 to include the dual-trace (switched beam) version, TF1331.

Because it contains many desirable features usually found only in much more expensive instruments, the Telequipment S.31 (previously the Serviscope) deserves mention. The specification includes: Y bandwidth, 0 to 6 Mc/s; maximum sensitivity, 100mm/V at all frequencies; rise time, 0·06 μsec; 9-step attenuator direct-reading in V/cm; 18-speed time base direct-reading in msec/cm and μsec/cm; × 10 X expansion symmetrical about centre of screen; automatic triggering up to about 1 Mc/s; trigger level control selects any point on signal slope for repetitive, random or one-shot triggering, by positive, negative, or TV line or frame; voltage calibrator; flat-faced c.r.t. The Y amplifier has two push-pull stages (the first a long-tailed pair) each working into a pair of cathode followers. The 2·7 kΩ anode loads are built out to 6 Mc/s by series inductors. Z.f. feedback is used to minimize astigmatism by keeping mean Y-plate potential constant. The time base is very similar to that described by Speight (p. 163)—a Blumlein integrator with something like a Puckle flyback system. Although intended primarily for television servicing and priced accordingly (£75) it is a good general-purpose easily portable oscilloscope. A double-beam version (D.31) is available, and also one with a higher performance (S.42). Model D.33 is similar in general specification, but alternative plug-in amplifiers enable it to cope with a range of applications for which heavier and much more expensive instruments have usually been required. One of the amplifiers gives the exceptional sensitivity of 0·1 mV/cm from 5c/s to 150kc/s.

Although the performance of all-transistor oscilloscopes does not compare with that of the best valve types, it is sometimes convenient to be independent of the mains. Microcell Type 165 has a 3½in. tube and the by no means contemptible Y-amplifier bandwidth of 0 to 10 Mc/s.

There are, of course, kits; notably a Heathkit and the Cossor 1045K, which are remarkably similar in external appearance, and both include printed circuits. Extremely full instructions are provided for making them up.

5.41. PHOTOGRAPHING OSCILLOGRAMS

Besides the obvious purpose of photography for making a permanent record of a trace—and making it easier to measure—it is needed for observing non-repetitive waveforms, either isolated transients or continuous but varying forms. Transients, and also very slow waveforms, can be made visible by a long-afterglow phosphor, or even better by means of the Memotron (p. 148), but these facilities may not be available.

The procedure with repetitive waveforms, which can be made to appear motionless on the screen, is straightforward: the camera is focused on the screen and an appropriate exposure given. For " one-shot " transient waveforms some device is needed to synchronize the shutter with the start of the sweep, or alternatively (if light can be sufficiently excluded) the shutter is opened for as long as is necessary to catch one sweep. To record continuously varying quantities the time base must be stopped (care being taken to avoid excessive brightness, which might " burn " the phosphor) and a long roll of film or paper drawn steadily along the X axis. Electric motor drives are obtainable for this purpose. Where still higher speeds are required, the film is mounted on a drum and rapidly rotated. Contacts are provided on the drum for starting and stopping the exposure.

Wherever possible, bromide paper is used for cheapness and convenience of handling, but in general film is needed for transients, and if they are high-speed a high e.h.t. is also needed. Blue phosphors, being the most actinic, are preferred for photography, especially at high c.r.o. writing speeds.

Many oscilloscopes are fitted for attaching the special cameras designed for the purpose, and available in considerable variety to enable one or more of the foregoing methods to be used. Footage indicators, synchronizing contacts, and drive motors are other features that can be obtained. The Langham Thompson Series 410 camera, for example, has a safe-light hood to allow the trace to be viewed during exposure, and a built-in developing unit, yielding prints 1 minute later. For occasional recording of " stills ", an ordinary camera with a close-up lens can be used in a not-too-bright room. A. J. Key (*W.W.*, June 1959) shows how to make an oscilloscope camera from a single lens and a few pieces of aluminium.

5.42. INEXPENSIVE ELECTRONIC INDICATORS

There are some purposes for which one would like to have the advantages of the cathode ray as an indicator, but where a quantitative indication is not essential. A diminutive and inexpensive tube is available for this purpose—none other than the " magic eye " tuning indicator. Being compact, cheap, inertialess, " unbumpable ", and negligible as a load, it may well be considered as a substitute for meters and other indicators. Certain types require only 3–4 V for the full range of movement, and give a visible response on quite a small fraction of a volt. In view of the very large input impedance, this represents a high degree of sensitiveness, which, of course, may be still further increased by amplification.

It is possible to make it sensitive to as little as 0·005 V z.f. by the simple expedient of connecting in series with the cathode a resistor of a few kilohms, preferably variable. This introduces positive feed-

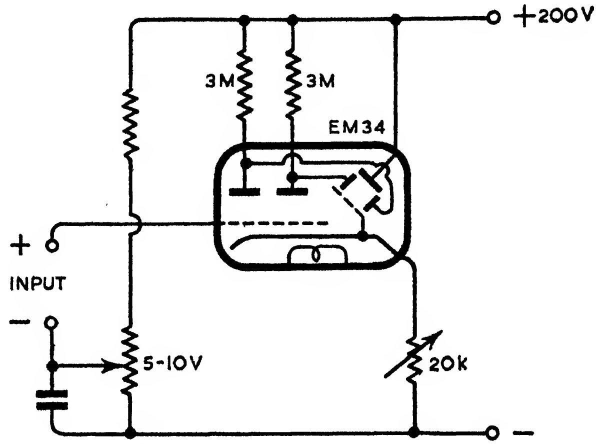


Fig. 5.52—Method of using the EM34 as a voltage indicator with controllable feedback to increase its sensitivity

back; but it does not work with all types of tube. The EM34 (which has the additional advantage of two " scales " with different sensitivities) is suitable, and Fig. 5.52 shows the makers' recommendation (*W.W.*, April 1947, p. 150).

Tuning indicators are popular as balance indicators in bridges (for example, Sec. 7.14), and are useful for indicating maximum in resonance methods. In slide-back methods they may even be used as substitutes for meters.

The Mullard DM160, being a sub-miniature measuring less than 1¼in. × ¼in., is of especial interest where space is limited.

Another interesting and useful type is the Philips E82M, with a push-pull deflection system, enabling voltages to be balanced (*E. & R.E.*, July 1957, p. 257).

Ordinary neon lamps are most useful in the laboratory as indicators and voltage stabilizers. It should be known that they are sold with or without a resistance in the cap: for laboratory purposes they are more useful without the resistance, but as the current then rises indefinitely at the striking voltage care should be taken to limit it in some way to a reasonable amount.

The photo-electric cell may be classed as an electronic indicator, responsive to light. At present it is finding application mainly in industrial fields, but although it is not a general-purpose laboratory instrument its existence should be borne in mind for possible special apparatus.

5.43. AUDIBLE INDICATORS

The indicating instruments described so far deliver their information through the eye, which is far more sensitive to changes of magnitude or form than the ear. They are therefore much better than acoustic indicators for such purposes as indicating maximum output from an amplifier. On the other hand, the ear is extremely sensitive to changes in frequency, and although the cathode-ray tube may be used for synchronizing frequencies with precision, so may phones or loud-speakers.

A pair of phones, especially of the adjustable type made by S. G. Brown, Ltd., is a remarkably sensitive instrument for detecting alternating currents of the order of 1,000 c/s. An audible signal is given by power of the order of 10^{-9} microwatt! So it is a usual choice for indicating balance in a.c. bridges working at the middle audio frequencies. The sensitivity both of ear and phones falls off so rapidly at the very low frequencies, however, that the same system is quite unsuitable for, say, 50 c/s.

Phones have the advantage of high sensitivity and of keeping out other noises and so improving concentration, but are tiring to wear for long periods and make one feel like a dog on a chain. So an amplifier and loudspeaker is sometimes preferable. An obsolete broadcast receiver, or one that is faulty on the r.f. side, may be used.

An important advantage of aural indication is that the ear can discriminate between signal and noise, or between fundamental and harmonic. And whatever progress is made in exact visual measurement, the ear remains the ultimate judge of much that our work is concerned with.

CHAPTER 6

Standards

6.1. PURPOSE AND BASIS OF STANDARDS

It was emphasized in Sec. 3.8 that in general it is better to measure a thing by direct comparison with a standard of the same sort than to deduce it from other quantities. Measuring a length, supposed to be 1 yard, by comparing it with a standard yardstick, is a sound and reliable method; measurement by observing the time taken by a standard snail to traverse the distance is likely to give the result to a substantially lower order of accuracy, for it is subject to various indeterminate influences. So among the most valuable equipment of a laboratory—particularly where accurate measurements are to be performed—are standards of the various quantities concerned. At the chief standardizing laboratory in this country—the N.P.L. (National Physical Laboratory, Teddington, Middlesex)—extreme precautions are taken to produce and maintain standards of the highest accuracy.

At one time the fundamental electrical standards of resistance and current were carefully defined material " yardsticks "—the resistance of a specified amount of mercury in a glass tube of specified dimensions, and the current that electroplated silver at a specified rate—and were the international ohm and ampere. But when it became possible to compare these more accurately with the absolute electrical units derived from the still more fundamental units of length, mass, and time, it was found that they were appreciably out. So from 1948 international units have been superseded by absolute units; and as standards of both sizes are likely to be still in circulation it is necessary in really accurate work to note that

1·00000	international	ohm	=	1·00049	absolute	ohms
„	„	ampere	=	0·99985	„	ampere
„	„	volt	=	1·00034	„	volts
„	„	henry	=	1·00049	„	henries
„	„	farad	=	0·99951	„	farad.

Although direct comparison with a standard is theoretically sound it is not always practically simple. However satisfactory our yardstick may be to check a length of ribbon, it might not be instantly obvious how best to use it for measuring the height of Mount Everest. In standardizing laboratories the standard of voltage is a special kind of cell. This rarely finds a place in a practical radio laboratory, which depends instead on indirect measurement by means of a voltmeter, calibrated by the makers from a sub-standard instrument, which in turn may have been compared with a standard cell. Although measurement by voltmeter is an indirect method (it not only does not compare the unknown with a standard voltage, but actually measures current!), the direct comparison method is inconvenient for ordinary purposes. It should perhaps be emphasized that most of what are called standards in this chapter would more properly be described as sub-standards, because they are instruments designed to retain with as great constancy as possible a calibration imparted to them from some primary standard.

The question of what the accuracy of an instrument means is, as mentioned in Sec. 3.6, not a simple one, and is discussed by W. H. F. Griffiths (*W.E.*, March 1943), with particular reference to standards of the quantities considered in the next few sections. For further information on the theory and construction of these standards, see B. Hague's *A.C. Bridge Methods* (Pitman) and L. Hartshorn's *R.F. Measurements* (Chapman & Hall). Standards for a.c. measurements were received by G. H. Rayner and A. Felton of the N. P. L. in *J.I.E.E.*, March 1961.

6.2. RESIDUALS

The first three standards to be considered are those of resistance, capacitance and inductance—R, C and L. The purpose of such standards is to provide an accurately known quantity of one of these parameters; preferably a wide range of values. To deserve the name of standard, or even sub-standard, the instrument must be so constructed as to retain its calibration against the ravages of time, temperature, humidity and reasonable use. Unfortunately it is not possible to provide any one of these parameters entirely without the others. The unwanted quantities are called *residuals*. So another important object in the design of a standard is to make the residuals negligible over the working range of frequency, or, failing that, to see that they can readily be allowed for. What one is entitled to regard as negligible depends of course on the degree of accuracy aimed at in any particular use of the instruments.

In an ordinary circuit diagram a single symbol is used to represent, say, a capacitor. But in considering standards at all carefully one must represent the residuals too. They cannot be completely represented by circuit symbols, because they are distributed throughout the component; but the real apparatus can generally be imitated well enough by quite a simple arrangement of lumped quantities—the equivalent circuit or network. For instance, a capacitor behaves very much as if it consisted of a pure capacitance shunted by a high resistance (to represent the leakage between the plates), both in series with a low resistance (to represent the resistance of the plates and leads and the loss in the dielectric) and a small inductance (to represent the inductance distributed throughout these conducting parts). If the imitation is a good one, the equivalence will hold over quite a wide range of frequency. But at any *one* frequency any network, however complicated, can be *exactly* represented by no more than two parameters—a resistance and a reactance (either capacitive or inductive). These can be either in series or parallel with one another. In this book they are denoted by R_s and X_s for the series pair and R_p and X_p for the parallel pair—though logically one ought to convert the parallel ones into conductance and susceptance (G and B). The reducing of more

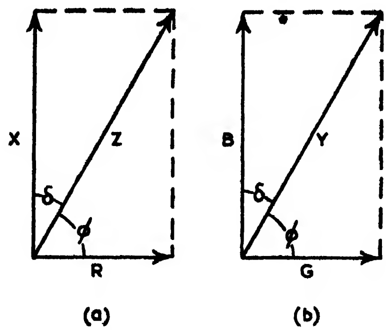


Fig. 6.1—Vectorial representation of (a) *resistance R, reactance X, and impedance Z, and* (b) *conductance G, susceptance B, and admittance Y, to illustrate the numerous alternative methods of specifying their phase relationship*

complicated systems to the equivalent pair is explained in Sec. 14.12.

The important bearing of this on any precise *R-C-L* work is that it is not enough just to say, for example, that the inductance is 165 μH. Which inductance? The inductance shown in the equivalent network? If so, which of the several recognized networks? Or is it the inductance corresponding to the reactive component in the reduced one-frequency equivalent? And, if so, is it the series or parallel value? Some methods of measurement give one and some another, so in comparing results it is necessary to be able to translate them to a common basis. All this sounds very alarming, but the less the impurities the more nearly equal are the various values of the main quantity. In a good standard capacitor, for example, the difference between C_p and C_s is negligible even in quite precise work. This is not always so with inductors, but a comforting fact is that unless there is reason for the contrary it is customary to work in series equivalents. The main thing to remember at this stage is that there *are* different values of any L, C or R, depending on where the residuals are assumed to be in the equivalent network. The question crops up not only in connection with standards but in the next chapter on bridges, and in Chapter 9 on measurement.

Some residuals, especially stray capacitance, depend not only on the apparatus itself but on its surroundings. If residuals are not small enough to neglect, then it is important that at least they should be definite and not liable to vary according to where the apparatus is used. That is why standards are usually screened—to ensure that all the capacitances concerned are always the same. This point is discussed in Secs. 6.6 and 7.6.

One more thing it may be as well to clear up before going on to the details of L, C, R, and their measurement: the variety of terms used to specify the ratio of resistance to reactance. It is easy to get confused with them all, especially as sometimes they are spoken of as if they were different and at other times as if they were the same. They can best be explained with a simple vector diagram, Fig. 6.1a, which shows how the series resistance and reactance, R and X, by which any network can be specified, add up vectorially to give its impedance, Z. The phase angle between Z and R is usually denoted by ϕ, and the " loss angle " between Z and X, which is $90° - \phi$, by δ. The parallel equivalent, made up of conductance G and susceptance B, add up similarly to give the admittance Y; the values of these in any particular circuit are of course different, but the angles are the same. The following are the things that are equal, whatever the values:

(i) $\cos\phi = \dfrac{R}{Z} = \dfrac{G}{Y} = \sin\delta =$ power factor

(ii) $\cot\phi = \dfrac{R}{X} = \dfrac{G}{B} = \tan\delta = \dfrac{1}{Q} =$ dissipation factor (or loss tangent)

(iii) $\tan\phi = \dfrac{X}{R} = \dfrac{B}{G} = \cot\delta = Q =$ magnification factor

When $R << X$, as it should be in a good capacitor or inductor, the first two lines are practically equal to one another and to the angle δ in radians, and are small.

Some writers, not content with this choice of alternatives, use another—the time constant. This is CR or L/R, depending on the kind of reactance. The advantage is that although C and L are not entirely independent of frequency they are approximately so over a wide range of frequency, whereas X obviously is not. In a resistor that is supposed to be non-reactive, the time constant ought to be a small fraction of a microsecond.

6.3. STANDARDS OF RESISTANCE

One kind of standard even the most modest laboratory can hardly do without is resistance. The most generally useful form is the decade box, consisting of a number of groups of ten equal-valued resistors (units, tens, hundreds, etc.) each group controlled by an 11-stud switch (0 to 10). For example, a three-bank box would provide a total of 1,110 Ω in steps of 1 Ω, or 11,100 Ω in steps of 10 Ω (Fig. 6.2). The usual type, made by the well-known instrument firms, has an inaccuracy within 0·1–0·2% and is reliable at all audio frequencies and well into the radio frequencies. For some purposes it is helpful to have a very wide range of resistance without unreasonable cost or bulk. The Dawe Type 201 box contains resistors from 100 Ω to 100 MΩ in the roughly logarithmic 1–2–5–10 sequence of values.

To preserve these valuable instruments from damage it is worth making a table of maximum permissible current for each range, and the

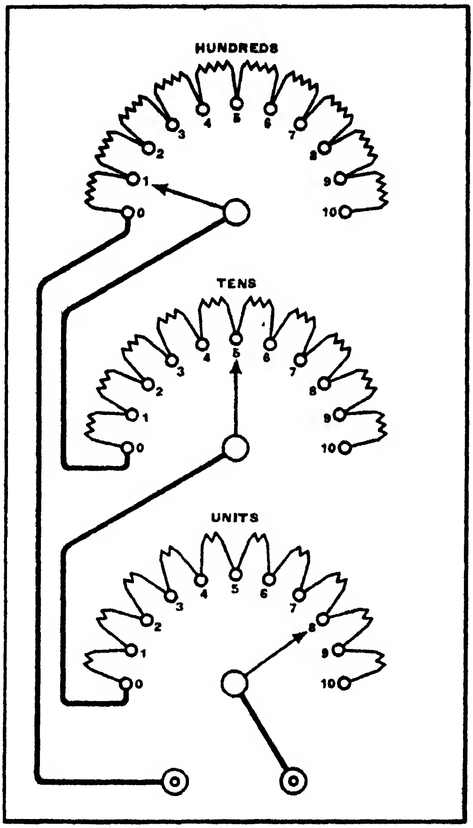


Fig. 6.2—Connections of a triple-decade resistance box, providing 1,110Ω in steps of 1Ω

corresponding voltages, to be pasted on the box; for it is very easy, in concentrating on an experiment, to exceed the limits and permanently impair the accuracy. Another way of preserving precision resistance boxes is to use less valuable ones when such high accuracy is not really needed. For example, there are Nash & Thompson "preferred value" resistance boxes, from 1·2Ω to 10 kΩ, and 1·2 kΩ to 10 MΩ. R. E. Thompson & Co. make combined *R* and *C* decade boxes at very reasonable prices. Separate ± 1% resistor and ± 5% capacitor boxes are produced by Winston Electronics.

"Bargain" resistance boxes ought to be accepted with caution, because their cheapness might be due to their belonging to the d.c. era, when an ohm was an ohm and nobody asked awkward questions about reactance. The plug connector system, once universal for resistance boxes, is extremely trying to anybody who has once used dial switches.

Whether incorporated in decade boxes or not, resistors to be relied upon to better than 1% have to be carefully constructed. For good-quality laboratory apparatus, manganin wire is favoured, because it has a high resistivity which is almost unaffected by temperature, and there is very little thermoelectric effect at a junction between it and copper. But it is rather tricky to use (Sec. 14.40). A more recent alloy is Karma,* made by the British Driver Harris Co. of Cheadle Heath, Stockport, from whom details are obtainable. It is similar to manganin as regards low temperature coefficient and thermoelectric e.m.f., but has nearly three times the resistivity and can be drawn into very fine wire, so much so (0·0006 in. dia) that 6·6 kΩ per yard is obtainable! This helps in reducing skin effect at high frequencies (Sec. 14.5). It is however more difficult to solder even than manganin. The American equivalent is Evanohm.

For making resistors oneself, Eureka (or Constantan) is more readily obtainable, is much easier to solder (an important point with very fine wires), has 29 times the resistance of copper, and for most purposes the effect of temperature is negligible—0·002% per °C (Sec. 14.5). But a copper-Eureka joint generates a very appreciable e.m.f. when heated, so when a soldered connection has been made it is essential to wait until it is quite cool before checking the resistance. And to avoid being misled by thermoelectric currents it is advisable to run Eureka resistors at a very conservative current rating.

It is not always realized that winding a resistor and adjusting it to within, say, 1 in several thousands does not produce a permanent standard of that order of accuracy. The process of winding inevitably sets up strains in the wire, and a resettling which goes on for a considerable time afterwards causes the resistance to change by an amount which may be as much as 1%. To avoid this, any resistors which are intended to remain highly accurate must be "aged" by annealing—baking them for up to 24 hours at 135°C. See also Sec. 14.40. Final calibration must be made after this process. One reason for keeping the subsequent working temperature moderate is to prevent change of resistance by further annealing. Excessive humidity, too, should be avoided; it can alter the resistance by 0·04%.

With low values it is important to consider the resistance of the switch contacts and leads. The method of use, wherever possible, should eliminate these from the calculation, by working in differences as explained in Sec. 3.9. But even so, these undesired resistances

* "The Electrical Characteristics of a Nickel-Chromium-Aluminium-Copper Resistance Wire", by D. Starr and T. P. Wang. *Proc. I.E.E.*, Part B, Sept. 1957.

should be minimized; in particular, a reliable switch should be used. Type 6001 switch by W. G. Pye has been specially designed for instrument work and is available with a variety of poles and ways. The overall switch resistance is about 0·7 mΩ and the variability ± 20 μΩ. Thermal e.m.f. is 0·025 μV/°C with temperature gradient across the switch. These properties have been retained after about 20 million operations.

For less precise purposes the comparatively cheap wafer types used in receivers are generally satisfactory.

The temperature coefficient of copper is about 0·4 %/°C, so its resistance changes very noticeably with ambient temperature, and still more when heated by current. Although nobody is likely to try to make a standard resistor of copper wire, there is the possibility of overlooking the fact that shunting a copper-wound milliammeter coil with Eureka is just about as bad. It is necessary for the moving coil to have a swamping resistance of low-temperature-coefficient wire in series with it, preferably many times the resistance of the coil itself.

6.4. NON-REACTIVE WINDING

Next to the accuracy and constancy of resistance, the most important thing is reactance, or rather the absence of it. Except for

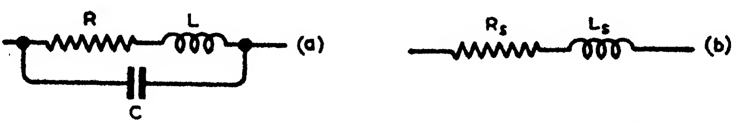


Fig. 6.3—Diagram (a) *represents with good approximation a resistor, L and C being the residual inductance and capacitance. At any one frequency, L and C can be represented as an equivalent series inductance L_s, as at* (b); *but the resistance R_s is in general different from R*

resistors incorporated in purely d.c. instruments such as moving-coil meters, it is hardly worth while buying or making resistors that are not wound so as to be usable over as wide a range of frequency as practicable. The equivalent shown in Fig. 6.3*a* represents any reasonably non-reactive resistor satisfactorily with fixed values of R, L and C, at all frequencies well below the resonance of L with C. Within this frequency limit, L_s in the simple series equivalent *b* is almost constant:

$$L_s \simeq L - R^2C$$

Since the effects of L and C are of opposite sign, cunning persons devise methods of winding by which they cancel one another out. E.g., if a 500-Ω resistor has an inductance of 2·5 μH its self-capacitance should be 10 pF. L_s is then zero. The apparent or measurable resistance, R_s, differs from the true value, R, as a result of the residuals, and within the same frequency limit

$$R_s \simeq R[1 + \omega^2C(2L - R^2C)]$$

So in this example it would be 0·1% high at 1 Mc/s. For minimum

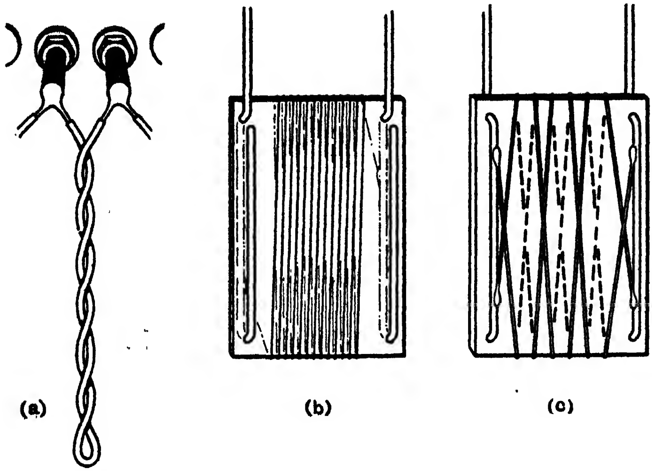


Fig. 6.4—Three of the many methods of minimizing reactance in resistance elements. The " twisted-hairpin " type (a) *is suitable for low resistance;* (b) *consists of fine wire wound on a thin mica card, which is stiffened by the thick wire connecting leads. In* (c) (*the Ayrton-Perry method*) *there are two windings in parallel, in opposite rotation to neutralize self-inductance*

variation in R_S one would aim at making R^2C equal to *twice* L, instead of L as for minimum L_S. L tends to be greater with large resistances than small, because of the greater length of wire; but relative to the resistance (which is what matters) it is more troublesome the lower the resistance. C, naturally, is more troublesome the higher the resistance. The easiest values of resistance to make accurately and non-reactively are the medium ones, say 10 to 1,000 Ω.

Low-resistance elements, of only a few inches of wire, can be made like twisted hairpins, as in Fig. 6.4*a*. Higher values, up to thousands of ohms, can be wound with very thin wire on the thinnest sheets of mica capable of supporting them firmly. The object of these constructions is to minimize the sectional area of the coil, and hence its inductance. The capacitance is quite low, too, if the wire is wound in a single layer from end to end along a small strip. Such construction, especially if with finer than about 40 s.w.g., is satisfactory at quite high r.f. and is sometimes adopted in signal generator attenuators. The wire must not be allowed to bulge out from the sides, but should lie flat on the strip, and it is a good thing to give it a coat or two of very thin shellac varnish. The current-carrying capacity is low; and for the medium resistances—1 to 100 or possibly 1,000 Ω—the Ayrton-Perry winding is more satisfactory. The procedure at first is the same as just described—a single layer from end to end on a strip of insulating material—but the resistance must be just double that required. Then another winding is put on, exactly similar except that it is in the opposite rotation. The inductances of these two windings are in opposition to each other, and therefore it is allowable for them to be supported on material of appreciable thickness, and hence for the wire to be of a substantial gauge. In addition to more care being needed in winding, this system is a little more complicated to adjust, because both wires have to be shortened equally.

When it comes to resistances of the order of a megohm, it becomes expensive to wind them of wire, and almost impossible if strict non-reactive methods are adopted. Generally such values are used as multipliers for d.c. voltmeters and may be wound anyhow; or possibly for a.c. of 50 c/s, in which case it is enough to wind on a multi-groove former, reversing every alternate section to give a fairly non-inductive result. But for high frequencies or for very high values one can hardly do other than use non-wire-wound resistors. Ordinary receiver resistors are not good enough for measuring apparatus, for, apart from their tolerances (they can always be made up to desired values somehow), they are usually affected considerably by temperature, voltage applied, and the lapse of time. But great advances have been made during recent years in the production of high-stability resistors, some of carbon and some of deposited metal or metal oxide films; and one need no longer look on them as fit for rough work only. They are seen as multipliers in some of the most reputable makes of voltmeter; and, as for high frequencies, Fig. 4.35 shows an example of use in a calibrated attenuator up to over 300 Mc/s. It is fortunate that r.f. attenuators usually have a resistance of the order of 70 Ω, because it is in that region that such resistors have minimum variation with frequency. Lower values increase with frequency; higher values fall—perhaps catastrophically, as for example from 100 kΩ at l.f. to 30 kΩ at 200 Mc/s.

At frequencies of more than a few Mc/s, or even lower, there is trouble not only with reactance but with an increase in resistance due to skin effect. Low values, say 1 to 100 Ω, are usually made of short lengths of wire sufficiently thin to keep the skin effect within tolerable limits. The gauges for a 1% rise are given in Table 14.8. However, these figures strictly apply only to straight wire. When it is wound into a coil the rise in resistance is increased by what is called proximity effect, due to the distribution of current in each turn being influenced by the magnetic field of adjacent turns. In general this is not calculable.

The inductance of resistors at r.f. is not usually negligible, but it can be excluded from the result by seeing that the length and shape of the circuit is kept the same for all readings, variation of resistance being obtained by substituting wires of different gauge and material.

The characteristics of resistors, with special reference to noise and to r.f., are discussed by G. W. A. Dummer in *W.W.*, June 1956.

6.5. ADJUSTMENT OF RESISTORS

A good deal of time and patience are required for making reasonably accurate resistors, but it is often useful to be able to do so when making up special apparatus. The approximate length of wire is calculated (Table 14.26), and slightly more than this taken for measurement. The length of wire between the terminals of the bridge or other measuring instrument is gradually reduced until correct; then it is bent at the edges of the terminals to mark the exact length, and enough extra for soldering to its own terminals. Remember to let the joint cool before checking the resistance. If it is too high, it can be reduced by building up the solder a little way along the wire; if too low, and the wire is thick, by filing. The file is used also in adjusting shunts and other low resistances made of strip metal. If the wire is very thin, it may be better to make the final adjustments on an auxiliary section in series wound with a thicker gauge. Thick wire resistors can be adjusted by a thin wire section in parallel.

6.6. STANDARDS OF CAPACITANCE

Standard capacitors are important not only for their known capacitance as such, but as pure forms of reactance. So absence of

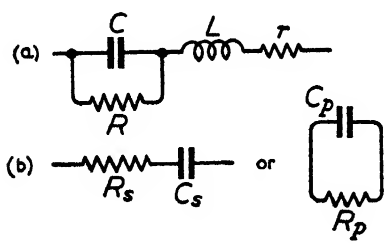


Fig. 6.5—To represent the behaviour of a capacitor over a range of frequency it is necessary to include parallel as well as series resistance (a). *At any one frequency, resistance and reactance can be represented either in series or parallel* (b). *Except at such high frequency that the reactance of L is appreciable compared with that of C, the differences between C, C_s and C_p in a good capacitor are very small*

resistance may be quite as important as correctness of capacitance. This side of the matter has been so well attended to by the leading instrument makers that in their best capacitors the series resistance is only a few millionths of the reactance. Another important thing at the higher frequencies is absence of the opposite kind of reactance—inductance. It is this that limits the frequency at which the capacitor is accurate. As explained in Sec. 6.2, the usually-accepted equivalent of a capacitor is Fig. 6.5*a*. Normally L and r are very small and R very large. At any one frequency r and R can be represented as a single series or parallel resistance (*b*) and L can be combined with C to give the apparent or measurable capacitance:

$$C_s \simeq C_p \simeq C(1 + \omega^2 CL).$$

So the greater C is, the more important it is to keep L small and the effect of L increases as the square of the frequency. R_s in any reasonably good capacitor is insufficient to make C_s and C_p differ appreciably. But that does not mean that no question need arise as to the value of a capacitor. The two conductors not only have capacitance between them but also to their surroundings, and unless properly screened these stray capacitances all vary according to the placing of the capacitor and its use. In Fig. 6.6, C_1 and C_2 represent capacitances to earth, and C_3 and C_4 to an unearthed body. Since changes of a hundredth of 1 pF can easily be detected with simple apparatus, these strays make a very noticeable difference, especially when C is very small; and one wonders what the calibration of a capacitor means.

In some methods of measurement (Sec. 9.16) all capacitances to earth are lumped in with the measuring apparatus, which reads only what is called the direct capacitance. In Fig. 6.6 this would involve C_3, C_4 and C_5 as well as C; so standard capacitors are provided with a screen which is normally earthed. This shuts off C_3 and C_4 altogether, and leaves only a fixed C_1 and C_2. Usually one terminal is earthed

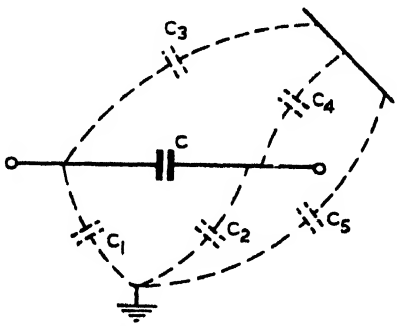


Fig. 6.6—The stray capacitances of an unscreened capacitor can be represented as shown, C_1 and C_2 being to earth and C_3 and C_4 to non-earthed surroundings. By the use of screening, these variable capacitances can be reduced to one fixed capacitance included in the calibration

too, say the right-hand one, thus shorting out C_2, so that the capacitance is $C + C_1$.

Even then, the unearthed terminal must poke out of the screen so that connection can be made with it; and what about its connecting lead? The most recent General Radio variable and fixed capacitors are provided with standard connectors to minimize errors here. They are also available in either two-terminal form (one terminal being earthed to case) or three-terminal, in which only the direct capacitance between terminals counts. The latter must be used in circuits where the relatively large and varying capacitances from terminals to earth do not affect the measurement. Standards of as little as 0·01 pF are available in this form, in which one plate " looks at " the other through a hole in an earthed screen.

However, the *difference* between any two readings on the scale of a variable capacitor is unaffected by whatever external capacitances may chance to be in circuit, so the difference method of reckoning (Sec. 3.9) is to be preferred, at least if 1 pF more or less matters in the reading.

Unlike resistance, capacitance is easier to vary smoothly and continuously than in small steps. So the most important laboratory types are variable air capacitors, seldom much more than 1,000 pF maximum. Isolated fixed capacitors are sometimes used, but more often a number are assembled in boxes and controlled by switches; in fact, one can obtain decade capacitance boxes uniform in appearance with resistance boxes; and very useful they are, if one can afford them. Where high accuracy is not essential, the price can be much lower, as noted in Sec. 6.3.

6.7. VARIABLE STANDARD CAPACITORS

The slightest movement, other than the desired rotation, of any of the parts of a variable capacitor is liable to upset the calibration very seriously. In the axial direction a shift equal to one inter-vane gap—which itself may be only a fraction of a millimetre—raises the capacitance to infinity! So the mechanical design and construction have to be of a very high order. Variable capacitors of the highest class are in fact among the most beautiful examples of scientific craftsmanship. Extreme rigidity as regards the constant dimensions is combined with smooth rotation. It is often assumed that a capacitor cannot easily be adjusted very precisely without a geared drive, but the gearing introduces mechanical problems of its own if it is not to cause errors. It is interesting to note that the extremely precise variables made by H. W. Sullivan, Ltd. have direct drive. Although offering considerable resistance to rapid rotation, the bearings are so beautifully made and adjusted as to give the feeling that if a fly leaned against the handle it would move steadily at an imperceptible rate. At least, there is no difficulty in making the most delicate adjustment by hand. Even the relatively cheap capacitors made by this firm are characterized by smoothness of rotation that surprises anybody whose experience is

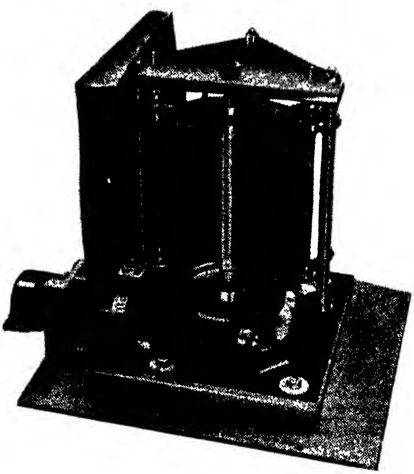


Fig. 6.7—Precision laboratory capacitor (Type D-14) designed to have an extremely small loss angle. The inner screen (shown mostly cut away) is connected to the rotor or low-potential terminal; there is a separate earthed screen, not shown except for the base plate. Note the earthed guard disks mounted between quartz insulators supporting the stator or high-potential system. (Muirhead & Co., Ltd.)

limited to ordinary "commercial" types. For information on the technique of making direct-reading scales to an accuracy of 0·01 %, see W. H. F. Griffiths (*W.W.*, Nov. and Dec. 1943). A more recent Sullivan development is a capacitor continuously variable over 100 pF (small enough to give a very open scale) and associated with ten steps

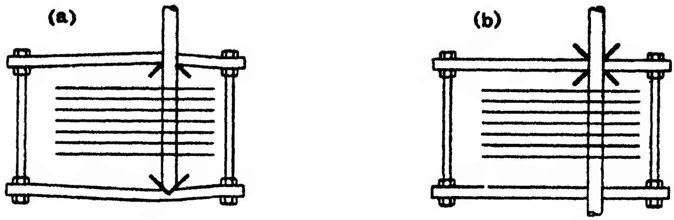


Fig. 6.8—Two types of variable capacitor bearing. Type (b), *in which the end plates are not under a separating stress, is to be preferred for permanence of calibration*

of 100 pF each, giving a total of 1,100 pF with the same low-loss characteristics as the variable air section.

In the General Radio, Muirhead and Marconi Instruments precision capacitors there is a gearing system in which the fast-speed drive is fitted with a separate scale for interpolating the divisions on the main scale. Naturally the mechanical precision of the gearing has to be exceptional. For the highest accuracy the errors of the gear can be measured and taken into account. For measurements on low-loss insulating materials, Muirhead have a double-screened model (Fig. 6.7) in which even the small loss due to the fused quartz insulators is excluded, and the vanes are mirror-polished, so the few remaining millionths of a radian of loss angle are due mainly to water vapour in the air. The capacitance is appreciably affected by the temperature and humidity of the air dielectric, and for long-term stability it would be necessary to seal the capacitor along with a desiccant. Design of capacitors for minimum temperature coefficient is discussed by D. A. Bell in *E.T.*, Sept. 1960.

Such instruments are not likely to be within the means of small laboratories. Even the home lab. ought to have at least one variable of sufficiently good construction to be worth accurate calibration. The vanes must have been treated to prevent warping, and rather more widely spaced than ordinarily. Bearings must give smooth rotation, freedom from side-play, undue wear, and slackening-off. The common practice of supporting the spindle by compressing it between the end plates is open to criticism, because even stout material subject to this stress is liable to acquire a set and so cause slackness and loss of calibration. It is better for the spindle to have free axial motion through one bearing, thus relieving both plates from stress, and for the other to determine the plate spacing (Fig. 6.8). Mechanical rigidity is achieved in the General Radio Type 1420 capacitors by milling them out of solid aluminium. Backlash or whip anywhere between the vanes and the scale must of course be negligible. A good geared drive is a great asset. Fig. 6.9 shows a 9:1 reduction type suitable for laboratory instruments. As regards vane shape, although fancy shapes—square-law, log, etc.—are used for special purposes, it is more usual for laboratory types to be of the old-fashioned straight-line-capacitance semi-circular shape.

The author finds it very useful to have a capacitor in which the fixed vanes are divided electrically into two or more groups, so that the range of capacitance is variable. In the lowest range, extra spacing or a smaller vane may be used to bring the variation down to perhaps 15 pF, for measurements such as valve capacitances. Two unequal

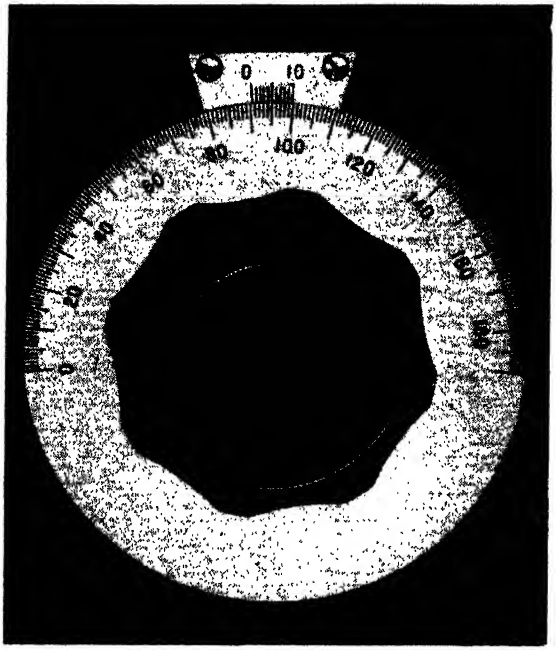


Fig. 6.9—Muirhead D-83-E slow-motion-drive dial with vernier. The scale plate is 4 in. in diameter. The drive, which can transmit a torque of 3 kg-cm, comprises a planetary reduction gear of the friction ball type. (*Muirhead & Co., Ltd.*)

groups of fixed vanes can be arranged by supporting them from opposite end plates; and this gives three ranges. The unused group, if any, is shorted to the moving vanes.

A simpler construction, substituting insulating fixed-vane supporting spindles and one set of insulating spacing washers, may be adopted if the somewhat higher losses can be tolerated.

The Sullivan capacitors are available in multi-range patterns, even down to the least expensive grades (Fig. 6.10).

For some purposes a very small variable is required—say 1 pF. Instruments in this range are available with micrometer drive (Fig. 7.20); or an actual micrometer can be adapted to provide an extemporized capacitor, by mounting it so that its moving plunger works inside an

Fig. 6.10—Dual-range variable air capacitor (Type C853A) with screening cover removed; maximum inaccuracy 0·03% (H. W. Sullivan, Ltd.)

insulated and screened cylinder. Or a miniature plate type can be made. They can be calibrated (preferably *in situ*) by a method such as that given in Sec. 9.34.

6.8. FIXED CAPACITORS

As it is mechanically impracticable to build continuously variable capacitors of large value, the range must be extended upwards by means of fixed capacitors. Complete variation up to, say, 1 μF, is obtainable by a three-switch decade box and a variable of 1,500 pF maximum (to give a working overlap). Such boxes, of first-class laboratory standard, are extremely expensive. Fortunately they are not in-

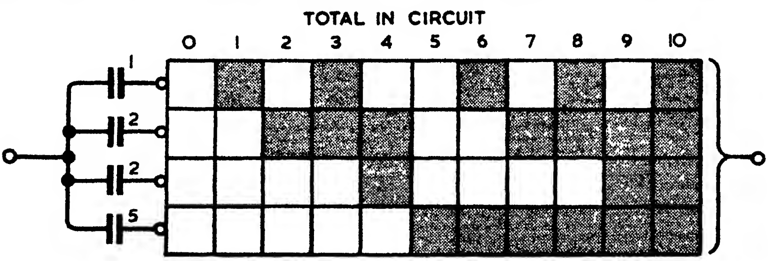


Fig. 6.11—Switching diagram for obtaining up to 10 capacitance units in steps of 1 unit, using four fixed capacitors. The shaded squares indicate switch contacts made

dispensable in a radio laboratory. But it is very useful to have a box giving at least a good approximation to values up to several microfarads. Even a box of rough standards is not so very cheap if it is on the straightforward decade plan, because special switches, progressively connecting the units in parallel, and a large number of units, are needed. However, a decade (0 to 10) can be obtained with four capacitors in the ratio 1–2–2–5, using a standard four-pole eleven-way switch connected as in Fig. 6.11. With a special Oak switch it can be done with

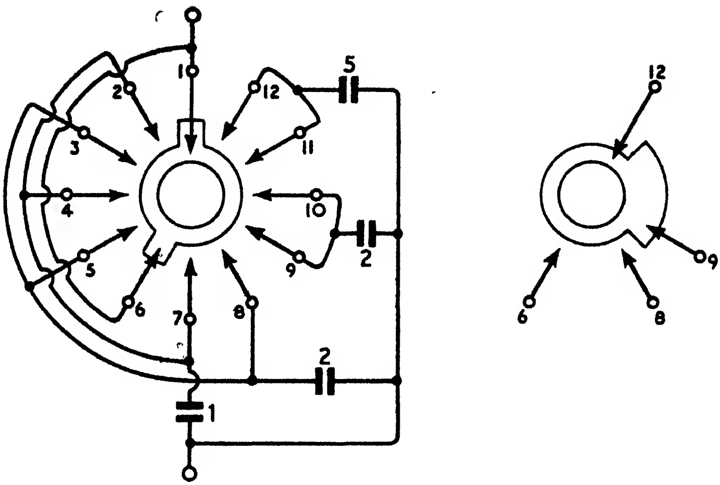


Fig. 6.12—Alternative decade capacitor switching using double-sided single-wafer switch. It is shown from the rear in the position corresponding to " O ". Each clockwise step (as seen from the front) introduces 1 unit of capacitance

one double-sided wafer (Fig. 6.12) as shown by C. D. Lindsay (*W.W.*, Oct. 1956). And if even this is too elaborate, a useful instrument is a single-pole eleven-way switch giving a choice of the following values: 1, 2, 5, 10, 20, 50, 100, 200, 500, 1,000 and 2,000 nF. Lastly, a triple decade box giving 0·111 μF in steps of 100 pF, using ± 1 % silvered mica capacitors, is obtainable in kit form at a moderate price (Heathkit DC1).

It must be remembered that the ordinary moulded-in broadcast-receiver types, even though they may be of mica, are not to be relied upon to maintain their originally measured capacitance, nor even to be particularly low-loss. It is a chastening thought that even precision capacitors at the N.P.L. have been found to drift as much as 350 parts in a million over a period of years; this for 0·01 μF, but the average drift for those of 0·1 μF and over was 60 and the greatest 120. Because some of the highest-quality capacitors are of mica it does not follow that use of this material is a guarantee. Those made by depositing metal on high-permittivity ceramic material are likely to be more constant than cheap mica types. Polystyrene is superseding mica for some purposes, as it is freer from polarization effects and has an even higher resistivity. . Most of the established manufacturers can supply types coming between the mass-produced sort and the best laboratory grades. Sullivan offer a range of high-permanence mica capacitors at quite moderate prices, so long as they are not required to be accurately adjusted to specified values. For moderate laboratory requirements a power-factor of 1 % is bad; 0·2 % is fair; and 0·05 % is good. Above

0·01 μF, mica capacitors are costly, and for ordinary purposes good-quality paper types are usually substituted.

For building up fixed capacitors to an exact required value, Johnson, Matthey & Co. supply silvered mica plates, including a " grid " type by means of which the final adjustment is made by scraping.

6.9. STANDARDS OF INDUCTANCE

Last of the " Big Three " is inductance, subdivided into self-inductance (L) and mutual inductance (M). Standards of self-inductance are handicapped by not being even approximately pure reactance—the resistive component cannot be made negligible. And their capacitance is more serious than the inductance of a standard capacitor. The usual simple equivalent is the same as in Fig. 6.3 except for the reversed relative magnitudes of L and R. The validity of this equivalent, even for high precision, is shown by L. Hartshorn and J. J. Denton (*Proc. I.E.E.*, Part B, July 1956).

Mutual inductance, on the other hand, can be provided with a phase angle as small as in good capacitors, with the advantage over capacitors that M can be continuously varied right down to zero and even negative values. Unfortunately stray capacitances limit these very good and useful qualities to fairly low audio frequencies. But for bridge measurements at not more than about 1 kc/s a mutual inductometer is the most versatile and accurate kind of standard. Without it, measurement of inductance usually involves comparison with capacitance; and at a.f. the impedances of tuning coils, for example, are so extremely small compared with that of a variable capacitor that it is difficult to get accurate results.

With any type of inductor one has to beware of the external field. It is liable to introduce unauthorized mutual inductance with other parts of the circuit. And even if the things it embraces are not parts of the circuit they increase the resistance, and if they are metal they either reduce or increase the inductance. So standard inductors, self or mutual, ought to have the minimum of metal in their construction, and certainly no iron or steel. Eureka screws are available, however, and are relatively harmless. Ordinary iron cores are never used, so a large number of turns—and hence high resistance—are inevitable in high-valued inductors. Even special dust cores, sometimes used, cause a slight change of inductance with current, which therefore must be strictly limited. Metal is needed for the coil itself, of course, and the thicker the gauge (to reduce its ordinary resistance) the greater the eddy-current loss therein. By using litz wire, consisting of a number of separately insulated strands, this loss can be reduced, especially at high audio and low radio frequencies; but it must be used both intelligently and carefully, because improper application and imperfect insulation and jointing may more than offset its advantage. It is also more easily damaged than solid wire.

Screening standard inductors is inadvisable, because it increases resistance and reduces inductance to an extent that varies with frequency. The only sure way to reduce the external field to a very small amount is to make the coil toroidal (Fig. 6.13). This is awkward for the amateur to wind but the General Radio Type 1481 standard inductors are of this form. Otherwise, coils that are not to couple must be kept well apart—say 10 times their diameter—or placed at right angles to one another, or both. Connection errors are discussed by J. F. Hersh (*G.R.E.*, Oct. 1960).

6.10. INDUCTOMETERS

The well-known Campbell variable mutual inductometer is shown in Fig. 6.14. The primary consists of a pair of fixed coils, with a small secondary swinging between them in a parallel plane to vary the mutual inductance from a small negative value to over 100 μH. The scale can easily be read to a fraction of 1 μH. Fixed secondary coils, in groups of ten, coupled to both primary coils, are brought out to decade switches extending the range to 11,110 μH. Other features can be used to extend the range both upward and downward. An inductometer is the nucleus of several types of bridge for self and mutual inductance and capacitance. Although usable only at a.f. it need not be regarded merely as an a.f. instrument, for the inductance of r.f. coils is usually measured at a.f. to render the effect of self-capacitance negligible. While such an instrument is expensive, it is not more so than the best standard capacitors, and is much more useful. It is surprising that cheaper types of inductometer,

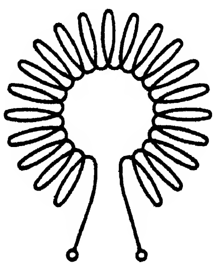


Fig. 6.13—Toroidal winding is an effective way of reducing the external field of inductors. A toroid is in effect a long solenoid bent round into a circle

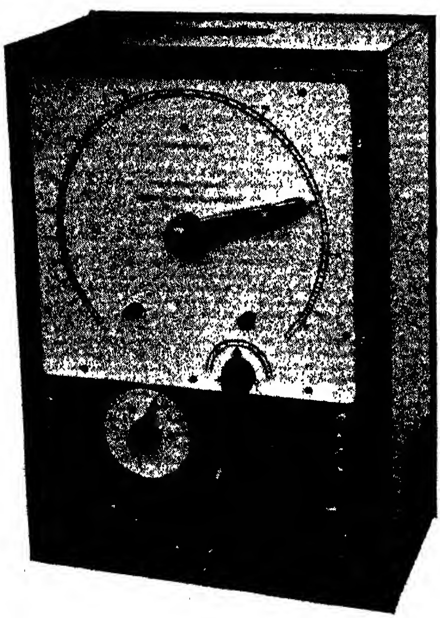


Fig. 6.14—One of the types of Campbell inductometer, for use in some of the most accurate methods of inductance measurement. Note the clear, open scale. (Cambridge Instrument Co., Ltd.)

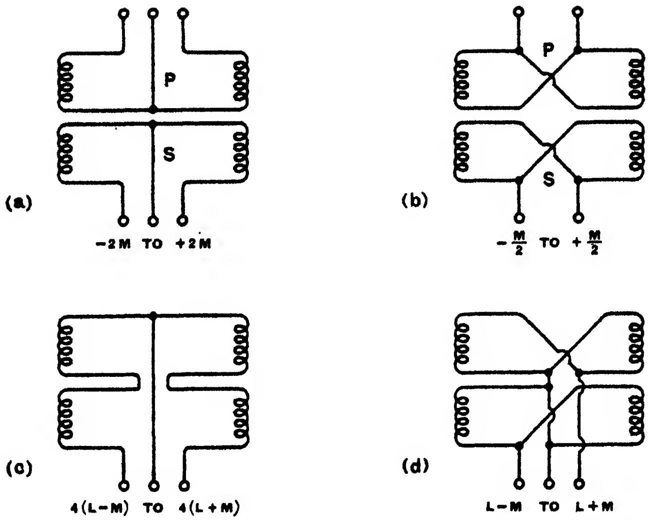


Fig. 6.15—Various methods of connecting a four-coil inductometer, giving approximately the mutual and self-inductances shown, where L is the self-inductance of one coil and M the maximum mutual between one pair of coils

retaining the same advantages in some degree, have not been put on the market, for the basic principle is simple enough. The chief difficulty, no doubt, is the time required to adjust to exact values.

Simple inductometers without tappings, for approximate measurements over a limited range, are quite easy. One method is to wind four identical coils, two fixed, and two moving on an arm pivoted midway, so that a 180° movement carries the coils from maximum coupling in one direction to maximum in the other. The two coils in each connected pair are wound in opposite rotation, which incidentally minimizes stray coupling to external coils. The coils can be connected in various ways (Fig. 6.15): (*a*) as a mutual inductometer, with balanced centre-tapped primary and secondary windings; (*b*) either or both pairs of coils in parallel for lower range; (*c*) as a self-inductance, all coils in series; (*d*) coils in parallel. If L is the inductance of any one coil by itself, and M is the mutual inductance between any two when they are closest, the respective ranges are: (*a*) $-2M$ to $+2M$; (*b*) $-\frac{1}{2}M$ to $+\frac{1}{2}M$; (*c*) $4(L-M)$ to $4(L+M)$; (*d*) $L-M$ to $L+M$. These are approximate, because they neglect coupling between coils that are not adjacent, but give some idea of the possible ranges.

M is inevitably less than L, and, if adequate clearances are allowed to prevent small unavoidable sideplay from causing large variations in the inductance, will be much less. So the ratio of maximum to minimum self-inductance is much smaller than in a variable capacitor.

However, a capacitor has nothing to compare with mutual inductance, which can be swung through zero to negative values.

The effect of mechanical imperfections on the calibration can be very greatly reduced by providing four primary coils, one on each side of each moving coil; this arrangement produces a nearly uniform field between them, and small axial displacements can take place without serious change in the mutual inductance.

6.11. FIXED INDUCTORS

A number of separate fixed coils of known inductance are useful in the laboratory, and as a set of them is probably required for r.f. measurements it is a good thing to have their inductances accurately measured so that they serve as standards. Obviously they should be as mechanically rigid as possible, and preferably with the lowest practicable r.f. resistance. Such a set of coils is almost essential in connection with a Q-meter and is obtainable as an accessory thereof (Sec. 7.20).

There are definite methods for designing coils with low r.f. resistance, and those who care to study them are referred to articles by R. G.

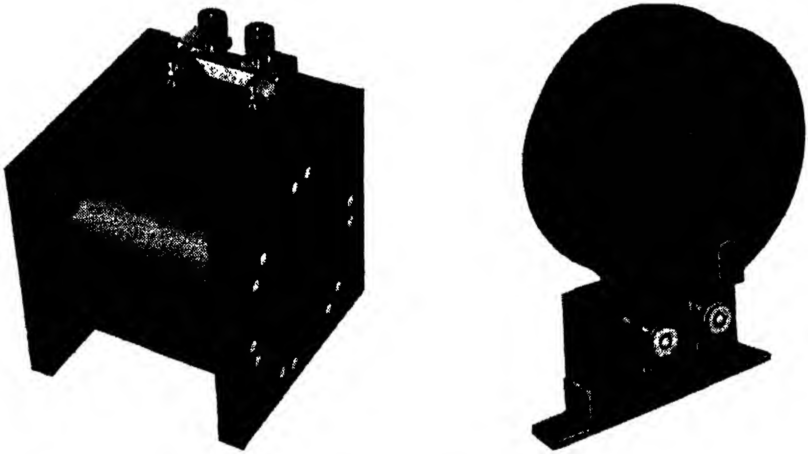


Fig. 6.16—Sullivan-Griffiths standard inductors, constructed on a compensating principle to ensure freedom from changes due to temperature. On the left is shown the first grade, for the highest precision. On the right is the second grade, giving excellent characteristics at low cost. (H. W. Sullivan, Ltd.)

Medhurst (*W.W.*, Feb. and March 1947) and a book, *The Theory and Design of Inductance Coils*, 2nd edn., by V. G. Welsby (Macdonald, 1960). One is not likely to go very badly wrong in using single-layer coils, wound with about one-wire spacing, on formers having a diameter 3 times the winding length, up to a few hundred μH; then multi-layer coils, so long as the layers are not too few and are properly spaced.

Standard coils in which the inductance is practically unaffected by temperature and age, made by Sullivan, have been designed on a compensating principle by W. H. F. Griffiths (*W.E.*, of Oct. 1929 and June 1934). They are marketed in two grades (Fig. 6.16), covering all requirements, from the home experimenter to the N.P.L. The accuracy of even the second grade is of the order of 0·1 %. Design of inductors for minimum temperature coefficient is discussed by D. A. Bell (*E.T.*, Sept. 1960).

6.12. STANDARDS OF VOLTAGE

As mentioned in Sec. 6.1, standard cells are not commonly used in radio laboratories, as they are expensive and must be looked after carefully. Comparison of the unknown is made by means of a

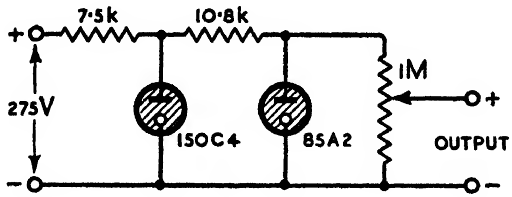


Fig. 6.17—Practical circuit for providing a precisely calibratable voltage, variable from zero up to about 85 V

potentiometer as explained in Sec. 10.2, so that no current is drawn from the cell. The e.m.f. of cells of the type used by the N.P.L., when a number are checked one against another, can be stable to $\pm$ 2 in 10^6 over a period of 20 years, so clearly they are among the most precise and reliable standards of any quantity at all.

The Mallory mercury reference-voltage type, obtainable from laboratory suppliers in single 1·35 V cells or 8-cell tapped batteries, though less precise is more practical and inexpensive. It is shock-proof, can be used in any position, and is capable of maintaining its voltage within close limits over a period of years, even when called upon to supply appreciable current, such as 100 μA.

Electronic voltage-stabilizing devices, notably neon tubes and Zener diodes (Sec. 4.3), if suitably selected and applied, can be relied upon to maintain constant voltages within very close limits. In this role they are sometimes referred to as voltage standards, but this is hardly correct, for they differ in two ways from standard cells. The voltage is not self-generated but has to be applied from a source, and it is not precisely known until it has been measured against a primary standard. So such devices are, at best, sub-standards. On the other hand, unlike standard cells they may be capable of maintaining their fixed voltage while supplying appreciable current.

Among neon tubes the Mullard 85A2 is an example; the voltage across a tube of this type may be anything from 83 to 87, but provided that the current through it is kept reasonably steady, preferably at about 6 mA, whatever the voltage for the particular specimen may be it remains constant within 0·5% throughout life, within 0·2% after the first 300 hours, and within 0·1% for any period not exceeding 100 hours after the first 300 hours. The temperature coefficient is only —0·003% per °C. The variation with current is about 0·35% per mA, so for the highest obtainable constancy it is necessary to stabilize the current, and Fig. 6.17 shows one method of doing so. In a circuit similar to

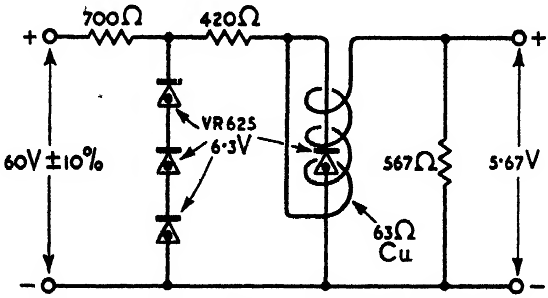


Fig. 6.18—Source of very constant low voltage, using Zener diodes as stabilizers, with thermal compensation coils around the output one

this the average increase in output voltage when the input was raised from 240 V to 300 V was found to be only 30 mV. In use, the output is balanced against the voltage to be measured, so that no current flows. Once an exactly-known potentiometer setting has been ascertained by balancing against a standard cell, the remainder can be calibrated directly in volts, since the open circuit voltage is proportional to potentiometer resistance below the tapping.

Zener diodes have the advantages of a much wider range of voltages (especially low values) and absence of extra striking voltage, and when properly used their constancy is comparable with that of standard cells; perhaps even better—see the design article by Chandler referred to in Sec. 4.3. Fig. 6.18 shows a typical circuit for output voltages up to 5 or so. Diodes with terminal voltages about 6–7 V have lowest and most easily compensated temperature coefficients, so are preferred for this purpose.

6.13. STANDARDS OF FREQUENCY

Frequency is undoubtedly the firmest foundation of all in the laboratory. It is unique in that one's own sub-standards can be checked at almost any time against world standards to within about 0·0000001%, which should be good enough for most people—and all at no charge for the information. Unique too is the precision with which one frequency can be compared with another, losing practically no accuracy in the process. It should not be forgotten that a standard of frequency is also a standard of time; the time interval between successive cycles is the reciprocal of the frequency. Constancy of frequency standards, based on atomic resonance, is now so great as to show up relatively large irregularities in the turning of the great earth itself. Clearly, then, an aim in devising laboratory measurements should be to obtain results in terms of frequency.

Details of broadcast standard-frequency transmissions are given in Sec. 14.37, and methods of frequency measurement and comparison in Secs. 10.9 to 10.15. In discussing oscillators (Chapter 4), frequency stability was kept prominently in view, because a reliable frequency calibration adds enormously to the value of any laboratory oscillator. So much progress has been made in this direction that well-designed signal generators can be regarded as sources of known frequency in addition to their primary function. It is desirable, however, to be able to check the calibration from time to time, especially when more than usual frequency accuracy is needed.

By processes of multiplication and division a single accurately-known frequency can be used to establish a succession of known points at as close intervals as one pleases. To cover the whole gamut of a.f. and r.f. in this way necessitates either fairly elaborate equipment or plenty of time and patience; for routine purposes it is more convenient to have continuously variable oscillators covering the required ranges. Even though their possible errors may be hundreds of times greater, their accuracy may still compare favourably with standards of capacitance, etc. And it is comparatively easy to check the calibration at one or more points by a really accurate standard, greatly reducing the risk of substantial error.

Frequency equipment can be summarized as follows, then. Where fairly rapid determination of frequency over a wide range with accuracy of the order of 1 in 10^8 is essential, even at high cost, the usual plant is a rack-mounted assembly containing a thermostatically-controlled quartz crystal of the highest grade, with a series of multivibrators and interpolating oscillators for relating its accurately-known frequency to any other frequency, and provision for checking against broadcast standards. This very refined equipment is too limited in appeal to come within the scope of this book, and information is readily available from the manufacturers and from the literature. But some account is given in the following pages of the simpler forms of frequency calibrator. Finally, at the higher frequencies especially, non-generating or absorption wavemeters are often very convenient.

6.14. QUARTZ CRYSTAL OSCILLATORS

The most commonly used frequency sub-standard is a plate or bar cut from a quartz crystal. Such a slice has a number of natural modes of mechanical vibration, the frequencies of which depend on shape, size and mode, and range from a few hundred c/s to 100 Mc/s. Quartz being piezo-electric, its vibrations generate corresponding differences of potential between opposite parts of the crystal, and vice versa; so by electrical feedback the vibration can be maintained continuously.

Although the vibration is mechanical, the piezo-electric effect renders the crystal equivalent to the electrical circuit of Fig. 6.19, in which the terminals represent the metal plates between which the crystal is supported —the only points of access to this " circuit ". The crystal is usually held at its nodes—lines or points of zero vibration amplitude. In some types, contacts are plated on to faces; in others, especially static installations where the maximum Q factor is wanted, a thin gap is left between crystal and plate. Q is also increased by evacuating the container, which is often a valve-type glass bulb and holder.

Although crystal-controlled oscillators are used as standards of very high accuracy, it should not be assumed that any oscillator which is

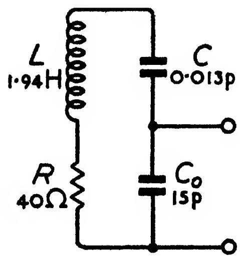


Fig. 6.19—Equivalent circuit of quartz crystal, showing typical values

crystal-controlled is necessarily a reliable standard. Much depends on how the crystal has been cut and how it is used. Inferior specimens have a nasty trick of jumping from one frequency to another near it, and the temperature coefficient is so high that unless the crystal is thermostatically controlled it may show little or no advantage over the best LC circuits. Then, as Fig. 6.19 shows, at one frequency it is a series tuned circuit (with a phenomenally high L/C ratio) shunted by a capacitance, and at another it is a parallel tuned circuit very loosely coupled by tapping down on the capacitance side (compare Fig. 4.17). Although these two resonant frequencies are very close together, they are obviously not identical, but confusion would be caused if a crystal calibrated by the maker in a series oscillator circuit were used in a parallel circuit or vice versa. In either case, the frequency depends appreciably on the circuit capacitance across the crystal—a fact which is very useful for fine adjustment, but if the frequency is to be as engraved on the case the capacitance has to be as specified by the maker.

Crystals vary considerably in their " activity ", i.e., readiness to oscillate, and instruments are available for measuring this property. From the data given in Fig. 6.19 (which apply to operation with a 0·6mm gap *in vacuo*) one finds that the Q of this specimen is 300,000. It would be substantially less in air and without a gap, and contamination with grease from handling, etc., would reduce it severely. It is only by reason of high Q that oscillation is possible in ordinary circuits with such unavoidably (and desirably) loose coupling.

Some oscillator circuits need conventional tuning as well as the crystal; Fig. 6.20, which relies on Miller-effect feedback from the tuned anode circuit, is an example. This reduces the risk of oscillation in an undesired mode, by confining effective feedback to a narrow range of frequency. In Colpitts-type circuits, oscillation is obtained without extra tuning circuits, as we shall see in Fig. 6.21. Low-activity crystals may do better in series circuits (H. B. Dent, *W.W.*, July 1952). Crystal activity can be defined by the largeness of the effective parallel resistance (R_p) or the smallness of the effective series resistance (R_s). If C_c is the effective capacitance of the circuit, in parallel with C_0 in Fig. 6.19:

$$R_p = \frac{1}{LC(C_0 + C_c)^2R}$$

$$R_s = R\left(1 + \frac{C_0}{C_c}\right)^2$$

Above about 20 Mc/s it is difficult to make a crystal oscillate in any fundamental mode, but by means of suitable mounting and circuit it may be persuaded to vibrate at a harmonic frequency—so-called overtone operation. This must be clearly distinguished from *electrical* frequency multiplication by selecting harmonics of the crystal frequency.

A bad crystal may have a temperature coefficient of the order of 100 parts in 10^6 per °C; the best, only about 1 in 10^6. For standards of

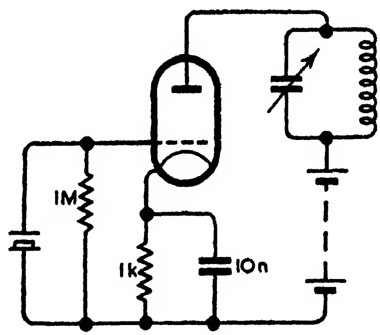


Fig. 6.20—Simple crystal oscillator for frequency checking

the highest precision, temperature control is needed anyway; but the ordinary lab. is likely to be interested in the best that can be done without going to the trouble. A good crystal is obviously the first necessity. That will ensure that the frequency will never differ widely from the nominal. For highly accurate determinations, one corrects the crystal by a broadcast standard, so what is needed is constancy during the time of a measurement. This is promoted by insulating the crystal from its surroundings, especially any heat-generating parts of the equipment. But it may not always be realized that the temperature of the crystal itself rises considerably if it is made to vibrate vigorously. It can even

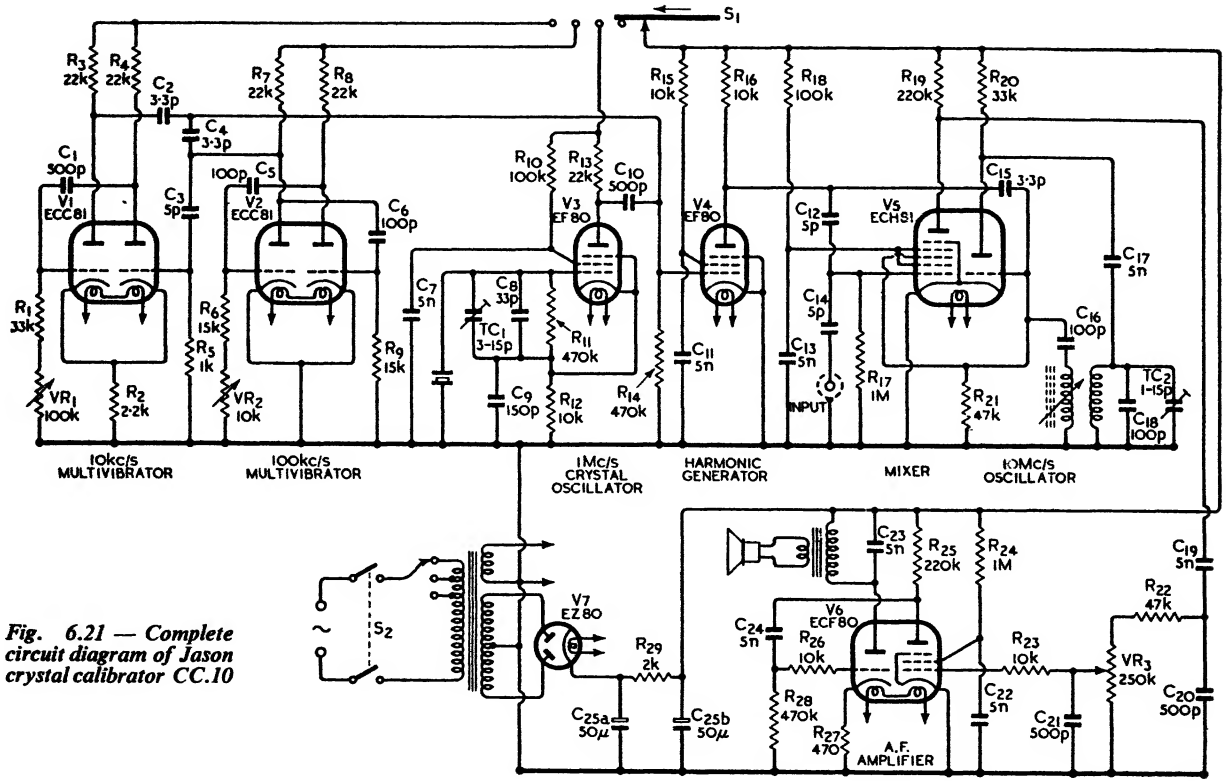


Fig. 6.21 — Complete circuit diagram of Jason crystal calibrator CC.10

be shattered or (short of that) permanently damaged. So it is wise to restrict its amplitude to the least that will do. The power into the crystal, which can be calculated as I^2R_c or E^2/R_c (where R_c is R_p or R_s as appropriate) should not be more than a few milliwatts at most.

A convenient alternative to the temperature-controlled oven for stabilizing crystal frequency has been developed by the Automatic Telephone & Electric Co. Ltd. (*W.W.*, Sept. 1958, and *E. & R. E.*, Aug. 1958). The essential is a thermistor, and in one example the frequency drift was restricted to about 2 in 10^6 from $-5°$ C to 45° C.

A comprehensive reference book is *Quartz Vibrators and their Applications*, by P. Vigoureux and C. F. Booth (H.M.S.O., 1950). A small practical guide to the specification and use of quartz crystals is published by the Electronic Engineering Association, 11 Green Street, London, W.1, at 5s.

6.15. EXTENDING FREQUENCY RANGE

For checking the frequency scales of signal generators, receivers, etc., one wants a signal of accurately known frequency at fairly frequent intervals over the whole vast range from a.f. to v.h.f. and perhaps beyond. The crystal oscillator itself inevitably generates harmonics, which can be encouraged in suitable ways to provide adequate signals at exact multiples of the crystal frequency, up to perhaps × 50 or more. If that is not enough, another oscillator can be locked to any one of these and a new starting point established. Sub-multiples of the crystal frequency can likewise be established by special types of oscillator. This could be avoided by using a sufficiently low-frequency crystal, but that is expensive. The commonest choice is 1 Mc/s or 0·1 Mc/s. It is obviously very convenient to have a nice round number like this. One should realize, however, that a crystal ground to a specified frequency within close limits costs a good deal more than one which may be every bit as constant (and that is the important thing) but a little odd in frequency. The latter can often be obtained quite cheaply second-hand. In any case, as we have seen, the frequency depends slightly on the external capacitance.

The problem in designing a crystal-controlled frequency calibrator (usually abbreviated to "crystal calibrator") is to provide ample spot frequencies for checking apparatus, but not so many that there is uncertainty in identifying them. Some kind of identifying and/or interpolation system is often included.* To distinguish the calibrator signals as a whole from other signals in a receiver with which it is being used, F. G. Rayer uses a neon tube to break them up into pulses (*W.W.*, Dec. 1960).

For maximum frequency stability the crystal oscillator stage itself should be run at low amplitude, with as little harmonic generation as possible. One of the most effective harmonic generators, which can easily be locked to the crystal oscillator, preferably through a buffer

* A practical self-interpolating crystal calibrator is described by D. Cooke (*E.E.*, Jan. 1952).

stage, is the multivibrator (Sec. 4.38). Besides being in this way a frequency multiplier it can also be used as a frequency divider, locking to every *n*th cycle of the primary oscillator.

The Jason CC.10 crystal calibrator (Fig. 6.21) is a good illustration, because it not only includes most of the features that have been mentioned but is obtainable inexpensively in kit form. V_3 is the crystal oscillator, of the Colpitts type, with TC_1 for adjusting the frequency to exactly 1 Mc/s. It is coupled through C_{10} and C_4 to a multivibrator V_2, which it holds in step at one tenth of its frequency. This multivibrator is of the symmetrical type shown in Fig. 4.46, in which the frequency is determined by C_6R_9 and by $C_5R_6 + VR_2$ as a fine control. In turn, via C_3, it holds another multivibrator V_1 to one tenth of its frequency. The second multivibrator is of the cathode-coupled type described in Sec. 4.38; its frequency being determined mainly by $C_1R_1 + VR_1$. Oscillations at all three frequencies, including many harmonics from V_1 and V_2, reach the grid of V_4, which by limiting amplitude develops harmonics, especially of 1 Mc/s, and passes them to both the control grids of V_5. Via C_{15}, the tenth harmonic from V_3 locks the triode *LC* oscillator to 10 Mc/s. This frequency, along with all the others, reaches the heptode section, together with the signal to be calibrated (via C_{14}). The output of this section goes to a two-stage a.f. amplifier, V_6. The 10 kc/s from V_1 is filtered out, so any a.f. note heard is due to the input signal beating with one of the internal fundamentals or harmonics. By bringing the oscillators successively into action by S_1, the multivibrator and *LC* oscillators can (with the aid of an approximately calibrated r.f. signal generator connected to INPUT) be set to the correct fractions and multiple of the crystal. At this stage the frequency uncertainty will be within about $\pm$ 0·01 %. Much closer adjustment can then be made by comparison with a broadcast standard, such as the B.B.C. 200-kc/s transmitter at Droitwich, the frequency of which is constant within 1 in 10^7.

When, as in this example, a frequency division by 10 is done in one step, care must be taken that the multivibrator has not slipped into the wrong submultiple. This should be checked from time to time. The design of decade multivibrators for reliable operation is dealt with by J. E. Attew (*W.W.*, March 1952).

Thrifty experimenters within range of Droitwich could—if they do not insist on having standard frequency laid on during the small hours—achieve phenomenal accuracy without the cost of a crystal by using the 200 kc/s carrier, suitably amplified, to control their multivibrators. Apparatus for doing this is available commercially—the Elliott Bros. Standard Frequency Receiver Unit Type 75D1547.

A simpler calibrator, battery-driven and portable, using a 1 Mc/s crystal and only two double triodes, is described by G. de Visme (*W.W.*, Dec. 1959). One valve works as crystal oscillator and a.f. modulator; the other as 100 kc/s multivibrator. Spot checks at these intervals are possible up to at least 30 Mc/s, and, by calibrating the lock control to divide the 1 Mc/s by numbers from 6 to 11 instead of only 10, other

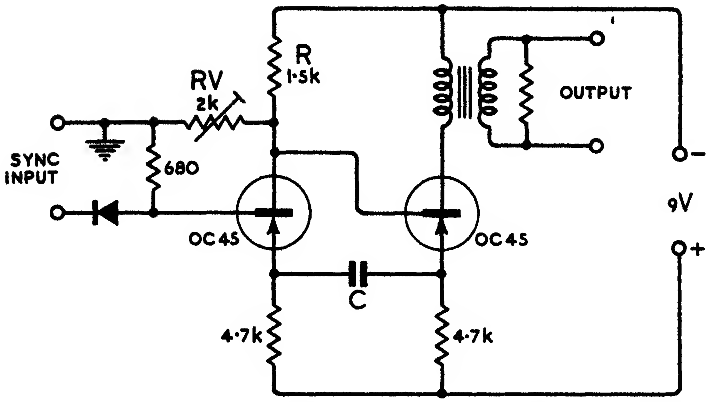


Fig. 6.22—Transistor multivibrator circuit suitable for frequency division. C and R are the main frequency-determining elements

controlled signals within a.f. beat of any frequency from 91 to 909 kc/s are available for lining up i.f. stages, etc.

An example of a professional frequency standard, using a 100 kc/s crystal in a controlled oven to give 100 kc/s and 1 Mc/s sine-wave outputs with short-term stability of 1 in 10^6–10^7, but lacking elaborate harmonic and comparison gear and so obtainable at a comparatively moderate price, is the Airmec Type 213. The same firm's Type 761 provides additional frequencies—10 kc/s, 1 k/c and 100 c/s—and includes a cathode-ray oscilloscope for Lissajous comparison (Sec. 10.10) and a synchronous clock driven from half the standard 100 c/s frequency, for time-signal checking.

The General Radio Type 1213-D (*G.R.E.*, Oct. 1959) claims a remarkably high frequency stability (short-term, of the order of 1 in 10^7) without temperature control, by using a crystal of such a cut and frequency (5 Mc/s) as to work on a nearly flat part of its frequency/temperature curve over the ambient range 0° to 40° C, and by protecting it from rapid changes by thermal insulation. The fundamental output frequencies are 10 and 1 Mc/s and 100 and 10 kc/s, with usable 10 Mc/s harmonics up to 1,000 Mc/s.

The foregoing crystal oscillator and frequency divider circuits have their transistor analogues, of course. The low impedance of transistors is most likely to be matched by the series resonance of crystals. The transistor multivibrator circuit shown in Fig. 6.22 appears to be particularly suitable for frequency division, among other things.* It is simple and " designable ", with a free-running frequency close to $1/3{\cdot}6CR$. RV is used to adjust its natural frequency to fall in with the frequency of the synchronizing pulses. On test this multivibrator was found to divide by 10 reliably over an ambient temperature range of

* " A New Linear Delay Circuit Based on an Emitter-Coupled Multivibrator ", by R. C. Bowes, *Proc. I.E.E.*, Vol. 106, Part B, Supp. 16, May 1959.

20° C to 60° C, and with 25 % changes in supply voltage. If the " flips " are synchronized, the " flops " should be arranged not to coincide with input pulses.

6.16. PASSIVE FREQUENCY STANDARDS

Although an oscillator is the most generally useful embodiment of known frequency, especially for work with receivers, and can be compared with other oscillators by beat-frequency and other methods, there are some purposes better served by the simpler non-generating or absorption types. Use is made of the very rapid changes in resistance and reactance of a high-Q circuit when it is tuned through resonance. The impedance of a series-resonant circuit, for instance, drops sharply to a very low value, so that it absorbs power from coupled circuits, and this can be read on some kind of indicator connected with either the coupled circuits or the resonant circuit itself.

The well-known absorption wavemeter consists simply of an isolated LC pair tuning over the desired band, sometimes with the addition of an indicator—a flash lamp or thermocouple in series (current-operated) or a neon tube or diode detector in parallel (voltage-operated). The lamp and neon tube give visual indication, and are useful in connection with senders and other oscillators that can provide the necessary power with the loose coupling needed for accuracy; the thermocouple is very easily burnt out; and both it and the diode necessitate suitable meters. An indicator on the wavemeter is not necessary if the oscillator being checked has a milliammeter reading anode current, merely tuning the wavemeter near the oscillator's tuning coil produces a flicker on the meter at resonance. A small tubular neon lamp in parallel with the wavemeter is the best idea, because it does not damp the wavemeter tuning when used below striking voltage with an indicator in the system under test, yet is ready to give its own indication from a powerful oscillator. The most accurate results are obtained when the coupling is reduced until the indication (neon glow or milliammeter kick) is only just perceptible.

Besides its simplicity and absence of feeding requirements, advantages of the absorption wavemeter are: fewer factors to upset the calibration; no wondering whether or not it has stopped oscillating; indication usually a simpler matter; no harmonics to cause confusion or uncertainty. It is particularly useful for the higher frequencies, more especially very high frequencies, at which it is difficult to " find the place " quickly with the heterodyne wavemeter when the frequency of the oscillator to be checked is quite unknown.

Apart from the obvious requirements that the coil or coils and variable capacitor should be constant and reliable and also low-loss, there is not much to be said about the design. To keep the operator's hands well out of the field of both wavemeter and " unknown " circuit it is usual for the tuner to be mounted at the end of a stick, with an extension control. At frequencies of hundreds of Mc/s (decimetre wavelengths) the absorption wavemeter may consist of a tiny capacitor with a little loop of wire connecting the terminals. Fig. 6.23 shows an example of this rigged up in a few moments, but adequate for rough comparisons; and also one covering all frequencies from 2 to 100 Mc/s with four coils. Note the comparatively large diameter of the coils, for easy coupling to the circuit under test.

If the oscillator to be tested is neither powerful enough to give an indication on this simple type of absorption wavemeter without excessively close coupling—or at all—nor is itself fitted with an indicator

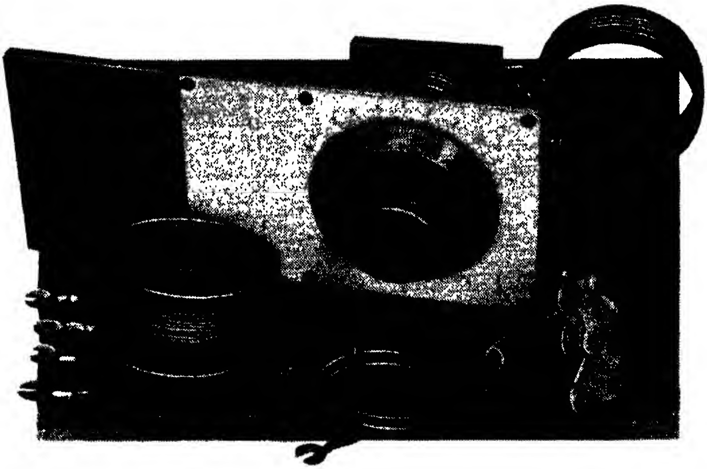


Fig. 6.23—Simple absorption wavemeters and coils; the larger covers 2 to 100 Mc/s and the smaller is for quick checking of decimetre waves

which could respond to the absorption, the sensitivity of the wavemeter needs to be increased by amplification. So as not to sacrifice the virtues of compactness and portability, transistors are preferred to valves. Their application here is not free from pitfalls, however, which have been explored by G. W. Short (*W.W.*, April 1959). In particular, it is only too easy to increase the response to relatively strong signals while not greatly improving the sensitivity to weak ones. Of several varieties, he recommends that in which a second transistor is used to modulate the r.f. input to the diode detector, so that the amplifier transistor works at a.f. and gives an audible as well as visible output. A 1½-V cell is sufficient power unit.

Although not a passive instrument, the grid-dip oscillator may be mentioned here because it is akin to the absorption wavemeter. It is simply a frequency-calibrated valve oscillator with a microammeter in the grid circuit. An alternative to the microammeter for monitoring grid rectified voltage is a " magic eye " tuning indicator. The instrument is used for finding the resonant frequencies of non-oscillating circuits. Some variations on this theme are provided by H. B. Dent (*W.W.*, March 1957). The Cossor Model 1461, with seven plug-in coils to cover 1·3 Mc/s to 300 Mc/s, can be used either as an absorption wavemeter or a grid-dip oscillator as required.

6.17. STANDARDS OF WAVELENGTH

Wavelength and frequency are fundamentally quite different quantities, as different as the length of a journey and the time spent on it; they are only interchangeable on the assumption that the speed of travel is the same in each case. While this is not even approximately true for journeys in general, it is true for electromagnetic waves in free space, and so a standard of wavelength is also a standard of frequency, and vice versa; either is calculable by dividing 300,000,000 (more correctly, 299,792,500) by the other, assuming wavelength to be in metres and frequency in cycles per second. But the assumption "in free space" must not be overlooked; in certain practical cases, to be referred to later, it does not apply.

Frequency is a more important quantity than wavelength in nearly all electrical communication work, because communication channels must be allocated on a frequency basis to allow for their sidebands. By the use of harmonics it is possible to calibrate in frequency well up into the v.h.f. bands; and L. Essen and A. C. Gordon-Smith have shown how to extend the principle to 10,000 Mc/s, using a 1-Mc/s crystal oscillator as the standard (*J.I.E.E.*, Part III, Dec. 1945). But at frequencies above about 30 Mc/s it is difficult to ignore wavelength altogether, because it appears as a directly-observed quantity in such things as aerials and feeders. And for approximate measurements with very simple apparatus, wavelength calibration is easier.

Use is made of the principles of standing waves along parallel wires. There are several variants of the method; the most compact is the quarter-wave resonator, which is only a little over 8 ft long for a 10-metre wavelength. For wavelengths of 1 metre and less the method is particularly convenient. A pair of parallel bare wires is strung across the room or mounted on a wooden frame. As the presence of material (other than air) between the wires is a source of error, it is best to do without any spacers, stretching the wires tightly between supports and taking very great care to get them exactly parallel. The farther apart, the less the probable error in parallelism, but the greater the distance that material—conducting or otherwise—must be kept away. About 2 in. separation is suitable for waves of the order of 5 metres, and 1 in. for shorter waves. Except for the insulators at the ends there should be nothing but air within at least several inches of the wires.

One pair of ends is left unconnected, and the other is short-circuited. If an oscillator coil is brought near the short-circuited end in such a position as to couple magnetically (see Fig. 6.24) there is the usual meter kick at resonance, which takes place when the oscillator wavelength is four times the length of wire from the short-circuit to the open end. It also takes place if it is four-thirds, four-fifths, four-sevenths, etc., of the wire length. Actually waves do not travel quite

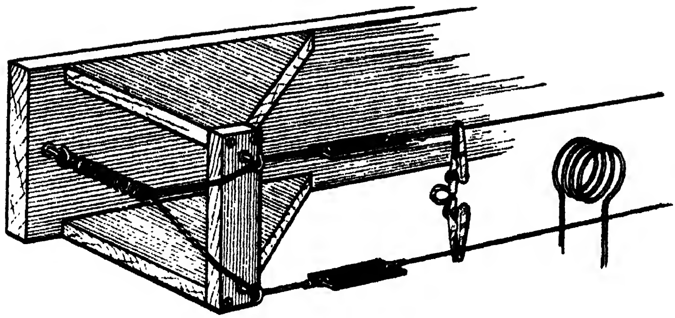


Fig. 6.24—Position of an oscillator coil for coupling to a parallel-wire resonator in wavelength measurement

so fast along such lines as in free space and the wavelengths thus measured are less by a small amount, but if the suggested precautions are taken the error should not be more than about 1 %.

Unless the oscillator frequency is approximately known beforehand, it may be difficult to find the indication of resonance within the tuning range of the oscillator. If it is found, it should be confirmed that it is due to the wire and not to an internal absorption, by bringing the oscillator away from the wire. Alternatively the oscillator frequency can be left alone, and the resonator tuned to it by shifting the bridge piece along. The awkward thing about this is that the oscillator coil must be shifted along with it. It is quicker to adopt the half-wave resonator by leaving the oscillator and the bridge at one end, and moving another bridge, made of a short piece of stiff wire pushed through the end of an insulating handle to form a T piece, away from it along the wire. At certain points the oscillator meter will probably jump about a bit; these points are half-wavelengths apart.

Further information on wavelength measurements, especially at higher frequencies—450 to 750 Mc/s—is given by H. B. Dent (*W. W.*, July 1958).

6.18. STANDARDS OF AMPLIFICATION: ATTENUATORS

It is possible, by such techniques as negative feedback and voltage stabilization, to construct amplifiers giving a known gain within fairly close limits. In fact, the previous chapter includes many examples of such amplifiers as necessary parts of indicators to read small currents and voltages. But amplification can be measured at least as conveniently by offsetting it against a known attenuation (Sec. 11.6) as by comparison with known amplification, and as a passive attenuator is simpler and more reliable than a standard amplifier it is clearly preferable. So a standard of amplification is, in practice, an attenuator.

6.19. THE POTENTIAL DIVIDER

If an attenuator works into a circuit that can be reckoned to be of infinite impedance—a voltage-operated device, such as a valve—

it can take the simple form of a potential divider or so-called potentiometer. The latter name is commonly but wrongly applied to the ordinary three-terminal rheostat or volume control, which, however, can earn some right to the name by being calibrated. But to be more definite the step-by-step type is preferred. The total resistance is made large enough not to be overheated by the current due to any

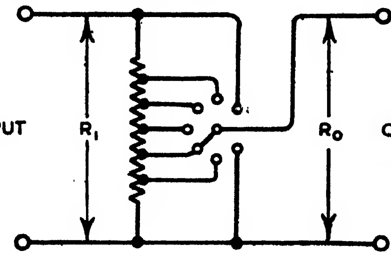


Fig. 6.25—Within certain restrictions, the simple potential divider can be used as an attenuator

voltage likely to exist across the input terminals, nor to load the input circuit excessively. With these conditions fulfilled, the lower the resistance the better, in order to minimize the effect of any finite impedance it may work into, and also to reduce the chance of interference from stray fields. On the assumption of infinite load impedance, the voltage attenuation is determined solely by the ratio R_0/R_1 (Fig. 6.25). It is often convenient to divide the resistance so as to give an equal number of decibels loss per stud. The value of R_0 for a given number of dB loss is given by multiplying the corresponding fractional voltage ratio in Table 14.17 by R_1. E. W. Berth-Jones has shown (*W.W.*, Feb. 1950) that provided moderate accuracy is sufficient it is possible to use preferred-value resistors, thus saving the expense or trouble of resistors specially adjusted to the odd values calculated as above. Fig. 6.26 is one of his examples—a potential-divider type of attenuator totalling just over 56 kΩ and giving 0 to 20 dB in steps of 2 dB. The errors shown are in addition to those due to resistor tolerances. These should therefore preferably be $\pm$ 1–2%, or wider-tolerance types adjusted by series or shunt resistors. The range of control can be extended to 40 dB by making up the lowest (5·6 kΩ) element with eleven resistors having the values given for the first 20 dB, divided by 10. Or the whole lot can be divided by 10 or 100 if a lower-resistance attenuator is wanted. *Multiplying* by 10 is not advised if stray capacitances are likely to be comparable in impedance at the highest working frequency.

The Dawe Type 202 is an accurate ($\pm$ 0·2%) wire-wound 10 kΩ potentiometer, and as it has three decades the ratio is adjustable from 0·001 to 1·000 in steps of 0·001. Each switch controls two decade resistances such that as resistance is added on one side of the output tap it is subtracted on the other. With capacitance loading less than

20 pF, the box is reasonably accurate up to 100 kc/s, and it will safely accept up to 230 V input.

6.20. LOADED ATTENUATORS

The potential divider is a particular case of what is sometimes called the L-type (or ⌉ or ⌈) attenuator, in which there are two arms as in Fig. 6.27, its peculiarity being that $R_1 + R_2$ is constant. Hence its attenuation can be varied by a simple tapping switch or slider. But it has the serious limitation that if the impedance into which it works (the load) is not infinite the attenuation is altered, and so is its input imped-

Fig. 6.26 (right) — Potential-divider attenuator utilizing preferred-value resistors. The 5·6-kΩ resistor at the foot can be replaced by a similar set of 11 resistors having one-tenth the values shown, to increase the range to 40 dB

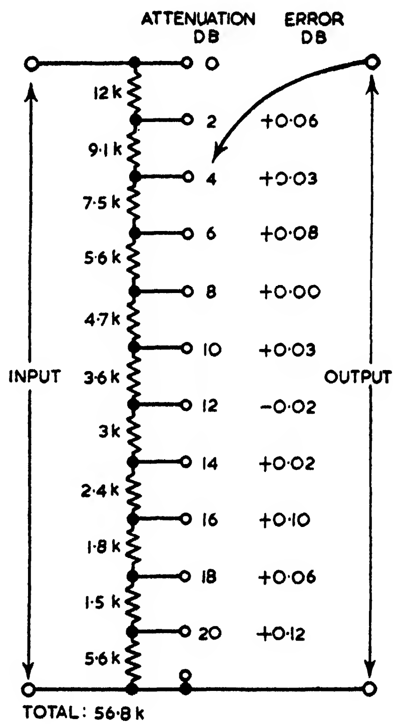
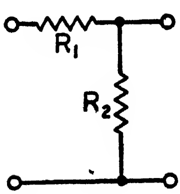


Fig. 6.27 (below)—The L-type attenuator, shown here, is the general form of attenuator of which the potential divider is a variety distinguished by $R_1 + R_2$ being constant



ance. To make the attenuation independent of load is clearly impracticable, so a loaded attenuator must be designed to feed a specified load impedance, and in practice that means a specified load resistance. To ensure constant source voltage it is desirable that inserting the attenuator or adjusting it should not alter the resistance seen by the source, i.e., the input resistance of the attenuator. And ideally this should not alter the resistance seen by the load (the output resistance) either.

If the attenuation is varied by transferring *conductance* from position R_1 to R_2, instead of resistance, decade division of voltage and current into a load can be accomplished. In one example there are nine 10–kΩ resistors, nine 100–kΩ resistors, and ten 1–MΩ resistors, which with the three decade switches all at minimum attenuation are all connected in parallel in position R_1 in Fig. 6.27, so $R_1 = 1$ kΩ. R_2 is then infinite, and the specified load (1 kΩ) connected across it receives one half of the input voltage. Each 10–kΩ resistor transferred from R_1 to R_2 reduces this ratio from $\frac{1}{2}$ to $\frac{1}{2} \times 0{\cdot}9$, $\frac{1}{2} \times 0{\cdot}8$, and so on. Similarly the 100–kΩ resistors provide 0·01 steps and the 1–MΩ resistors 0·001 steps. The input resistance obviously rises meanwhile, which may affect the input voltage, but the input/output ratio remains as shown on the dials. The output resistance also varies with switch settings, unless the source impedance is negligible. In an instrument of this kind designed by E. R. Wigan (*E.E.*, Sept. 1960) the switches also control a shunt across the input terminals in such a way as to keep the input resistance constant at 1 kΩ.

With this addition it amounts to a three-arm attenuator of the Π type one of the two basic configurations (the other is the T) which can be designed so that when inserted between source and load each continues to see the same resistance as before. These two resistances are not necessarily equal, but we shall consider only those that are. In Fig. 14.27 these two forms are shown, together with three others developed from them, with formulae for calculating the values of the resistors needed for any required input/output voltage (or current) ratio. Table 14.10 shows the values required to give the listed amounts of attenuation in dB when working between resistances of 1 ohm; when working between resistances of R ohms, multiply all r values by R.

The attenuation can be controlled either by simultaneously varying all the resistances by ganged switches or by using a set of change-over

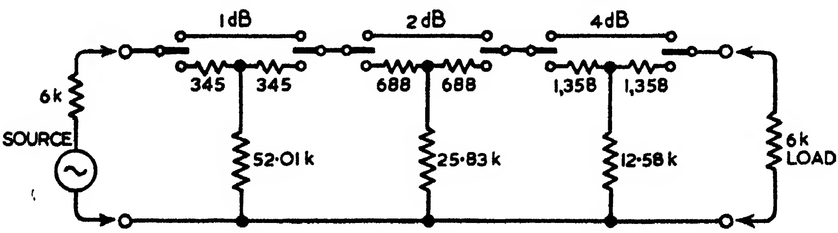


Fig. 6.28—Worked-out example of a three-stage T attenuator giving a maximum of 7 dB in steps of 1 dB

key switches to cut in, or out, the desired number of stages of fixed attenuation; the latter method is slightly less rapid to manipulate but is easier to construct and more suitable for high-frequency use because sections can be separately screened and points of widely different potential kept apart.

To illustrate this a three-stage attenuator is shown in Fig. 6.28. When all the switches are in the upper position there is obviously no attenuation, and the resistance of the far end, looked at from either end, is 6 kΩ. When the left-hand pair of switch arms—which form one switch—is depressed, there is no change in the resistance, but 1 dB attenuation is introduced. Similarly with the other stages. A total of 7 dB, variable in steps of 1 dB, can be inserted. Using seven separately screened sections in this way, up to 127 dB can be obtained.

The bridged-T type has the practical advantage that only two resistances, R_5 and R_6 (Fig. 14.27) have to be adjusted for each step. The other two are permanently connected and equal to the input and output resistance R. The switching can be done by a two-pole wafer type.

6.21. LADDER ATTENUATORS

For systems which must be balanced with respect to earth, the more complex H attenuator is used, and requires a set of four-pole switches. In radio equipment, for both a.f. and r.f., something more simple, compact, and quickly manipulated is wanted, and, although it does not maintain perfectly constant resistance with the switch near one of its ends, the ladder attenuator (Figs. 6.29 and 4.35) is much used,

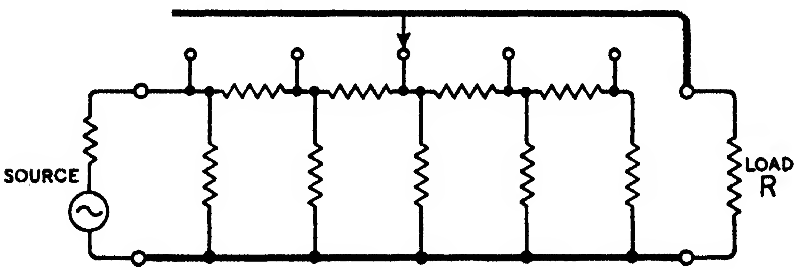


Fig. 6.29—Ladder form of attenuator, commonly used in r.f. signal generators

especially in signal generators and audio amplifiers. It can be regarded as developed from a series of Π sections, as in Fig. 6.30*a*. The attenuations of the sections are not necessarily equal, but this type of attenuator is usually required to give equal steps, so for simplicity assume the sections are identical. They can be designed in the usual way to work between equal resistances R_0 as shown, the formulae being

$$S = R_0 \frac{a^2 - 1}{2a} \qquad P = R_0 \frac{a + 1}{a - 1}$$

in which a is the input/output voltage ratio per section. The resistance between the terminals *tt*, when the termination R_0 is connected to the other end of the section, is therefore R_0; and similarly for any number of sections, looking either right or left. When they are all permanently joined together as at (b), pairs of resistances P can obviously be replaced by single ones $P/2$.

The ladder differs from the basic multi-section Π type in that the load is not the same thing as the termination but (as it were) gate-

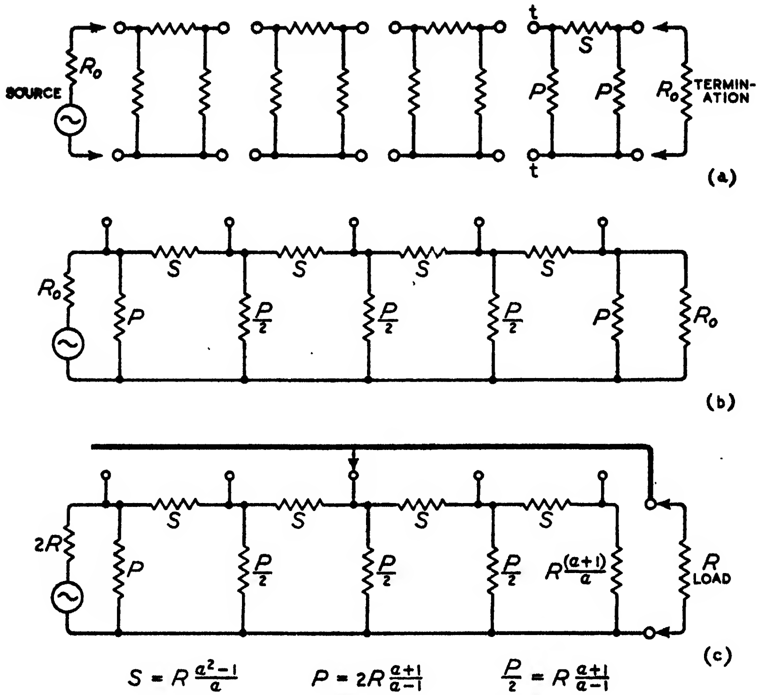


Fig. 6.30—Showing how the ladder attenuator of Fig. 6.29, repeated here at c, *can be regarded as a development of the* Π *type*

crashes into an already complete and matched party. To whichever stud it is connected (as in Fig. 6.29), it finds two resistances R_0 in parallel, one to the left and the other to the right, making $R_0/2$. If this output resistance is to match its own (R), then R_0 must have been made equal to $2R$. Substituting these values, and merging the termination with the last P, we get (c), in which P and S have the values shown.

Although the load matches the network as a whole between the points to which it is connected, the source and effective part of the attenuator (output resistance $2R$) are mismatched by looking into the tail of the attenuator *in parallel with the load*, i.e., $2R$ in parallel with R, which is $2R/3$. The total power received is therefore less than with perfect matching, and one third of it is wasted in the tail of the attenuator. The combined power loss, as compared with what would be received by a matching load $2R$ connected in the regular manner, works out at 50% or 3 dB. However, this loss is the same at every point of the switch, so the changes of attenuation between points are as designed. The resistance seen by the load is also the same at every point, and equal to R, at least in theory. In practice the source is an oscillator, and its impedance is likely to be a resistance $2R$ at all frequencies. If the load is switched directly across it, the impedance the load looks into may thus be appreciably unequal to R. But at other points the influence of the source impedance, being separated by attenuator sections, becomes negligible.

There is a continuously variable variety of the ladder attenuator, in which all the resistors S together take the form of the circular element of a " potentiometer "; e.g., Fig. 12.6. Ideally, the parallel elements should then also be merged into a uniformly distributed resistance, but matching is reasonably well preserved if they are tapped across at intervals of one or two dB.

At very high frequencies it is necessary to guard against reactances that would increase or reduce the nominal attenuation. Besides the usual methods for standard resistances, screening must be used between points at widely different power levels. For further information, see *M.I.*, Sept. 1954.

6.22. OTHER ATTENUATORS

For more detailed treatment of attenuators, including the effects of incorrect, unequal or reactive terminal impedances, an advanced work such as A. T. Starr's *Electric Circuits and Wave Filters* (Pitman) should be referred to. And there is a very valuable paper by McElroy in *Proc. I.R.E.*, March 1935, giving more extended data than in Sec. 14.24. It is possible, for example, to match unequal input and output resistances by an attenuator, but there is then a minimum insertion loss; the greater the inequality the greater this loss.

Quite a different type is the waveguide attenuator. It was devised principally for use at frequencies of thousands of Mc/s, for which attenuators using circuit elements such as resistors are unsuitable. One of the features of a waveguide as a means of conveying power is that above a certain critical frequency it is very efficient. That is to say, its attenuation is very small. Below that frequency the attenuation is very rapid, and the number of decibels is proportional to the length. So all that is required is a length of metal tubing, sufficiently narrow for the critical frequency to be well above the working frequency, and suitable electrodes at each end.

If the highest working frequency is not more than about one-seventh of the critical frequency, the attenuation does not depend more than 1 % on frequency, and is k/d dB per cm length, where d is the diameter of the tube in cm and k is 41·8 for concentric plate electrodes (E_{01} waves) and 32·0 for loops (H_{11} waves).

The disadvantage of this attenuator, which explains why it is not generally used for low radio frequencies, is that when the electrodes are close together the law is complicated, and it cannot conveniently be adjusted to less than about 20 dB.

CHAPTER 7

Composite Apparatus

7.1. WHAT THIS CHAPTER INCLUDES

For most investigations one uses combinations of the instruments discussed in the last three chapters, assembled and connected up for the purpose. Except for specialized work it would not be economic to have them made up permanently into single pieces of equipment. There are however some combinations so useful that they form distinct classes of their own, and these are the subject of this chapter. Bridges are the most important class. Although some are now made for use up to hundreds of Mc/s, bridges are more often used at frequencies lower than r.f. For high frequencies there is another important class of instrument, typified by the Q meter, employing resonance methods. Valve and transistor testing is often sufficiently important to justify a special equipment for the purpose. A number of other composite instruments, such as distortion meters, modulation meters, frequency meters, phase meters, etc., appear in manufacturers' catalogues, but seldom in the not-very-large general lab., so here they are described briefly, if at all.

7.2. BRIDGES IN GENERAL

Bridges are mainly for measuring R, C and L, either as such or as impedance or reactance; there are also bridges for measuring frequency and for valve and transistor parameters. One of the advantages of bridges is that they are null systems; i.e., the indicator has to do no more than indicate zero, which enables measurements to be made with high precision (Sec. 3.10). And as the results are in terms of passive standards they can be made with accuracy.

There is another, and for general purposes even more attractive, feature of bridges. The quantities quite ordinarily measured in a radio laboratory cover an enormous range. Taking capacitance: there is 0·001 pF as one extreme (the anode-grid capacitance of a r.f. valve), and 1,000 μF as the other (d.c. blocking in a transistor circuit). These enclose a range of a British billion to one (10^{12} : 1). Resistance is encountered over a rather wider range; inductance rather less. The idea of having to provide accurate standards for direct comparison covering all of these enormous ranges is alarming to contemplate.

That is where bridges come in. It is not practicable to measure the diameter of the sun against a standard yardstick with a pair of outside calipers, but by holding the yardstick a little distance away from the eye the sun can be measured if one knows the ratio of the distances from the eye to the stick and to the sun. A bridge is, firstly, a convenient system for comparing two electrical quantities with great precision, and secondly a means of introducing a multiplying ratio into the comparison. Comparisons can be further extended to different kinds, such as measuring L against a standard of C or R. The Wayne Kerr B.521 bridge, for example, which contains everything in a box about the size of a table radio set, covers 0·001 Ω to 1,000 MΩ, 1 pF to 5 F and 1 μH to 500 kH. Standards of equal accuracy, continuously variable between those limits, if practicable at all, would be excessively costly and occupy the whole laboratory.

The original Wheatstone bridge for measuring resistance with d.c. (Fig. 7.1), from which so many varieties have evolved, is a symmetrical network of six " arms ", one containing a source of current such as a

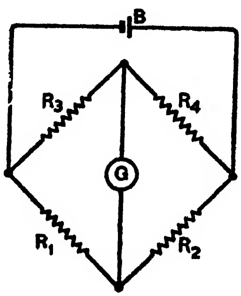


Fig. 7.1 — Simple Wheatstone bridge, on which all other forms of bridge are based

battery, and another a detector such as a galvanometer. When the resistances of the remaining four are in proportion, i.e.,

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

(which of course is the same as $R_1R_4 = R_2R_3$ and $R_1 = R_2\ R_3/R_4$) then the two points between which the detector is connected are at the same potential, so the indication is zero. The bridge is balanced. If any three of the resistances are known, the fourth can be calculated. It is not necessary to know the actual values of three; the value of one and the ratio of two are enough. So R_3 and R_4 (say) are called the *ratio arms*. If R_1 is the unknown, to be measured, then R_2 should be a variable standard resistance. Supposing first that $R_3 = R_4$; then at balance $R_1 = R_2$. By varying the ratio $R_3 : R_4$, measurements can be made beyond the range of R_2.

In theory there is no limit to the range of bridge measurement, but in practice the sensitiveness of indication, which is greatest when the resistances of all six arms are equal, falls off the more unequal they are.

The bridge network being symmetrical, the condition for balance is not affected by interchanging the source and detector. The sensitivity may be substantially affected, however; especially if the galvanometer resistance is high and the ratio large.

The same balance condition holds with a.c., but it must be interpreted more generally: instead of $R_1/R_2 = R_3/R_4$ it is $\mathbf{Z}_1/\mathbf{Z}_2 = \mathbf{Z}_3/\mathbf{Z}_4$, the bold type signifying that these are vector quantities, in which phase as well as magnitude must be taken into account. To get a balance, in other words, the potentials at the detector terminals must not only be equal in voltage but also in phase. The bridge must be balanced for reactance as well as for resistance. In practice it is unlikely that it would balance for both simultaneously unless special provision were made for adjusting these quantities independently. In certain types of bridge, more especially those including self-inductance, these adjustments are not independent, and one may work away at them, first one and then the other, for a considerable while before reaching a balance. Avoidance of this is an important aim when choosing a bridge.

If all four arms are devoid of reactance there is no complication, and with a.c. source and detector the bridge shown in Fig. 7.1 can be used for measuring resistance. In the radio laboratory an exclusively d.c. bridge is not usually justifiable. But measurement of resistance by a.c. is not always a satisfactory substitute for d.c., because a.c. resistance often differs appreciably or even largely from the d.c. value, and it cannot be measured at all with a resistance bridge if it is shunted by capacitance or is wound inductively. What is often done, therefore, is to provide for switching over to d.c. source and detector.

There are so many varieties of a.c. bridge that it is perplexing to decide which to use. The chief difficulty is choosing between bridges that can do a limited range of measurements very well and those that do almost everything not very well. What follows is an attempt to bring out the most important practical points concerning the types most likely to be useful in the unspecialized laboratory. Readers who want to know more, especially about the high-precision specialized types, should turn to Hague's *A.C. Bridge Methods* (Pitman). There is also an excellent chapter on bridges in F. K. Harris's *Electrical Measurements* (Chapman & Hall).

7.3. COMPONENTS OF BRIDGES

Source.—For d.c. the source is usually a battery, though there is no reason why a mains unit should not be used. In either case it is advisable to have some form of control, such as a rheostat in series, to limit the current through the bridge to a safe value. For a.c. the source is a fixed- or variable-frequency oscillator, such as described in Chapter 4. Although a variable frequency is necessary for some purposes, a fixed frequency is sufficient for the majority of measurements, and 1,000 c/s is a common choice. In calculations it is ω, equal to $2\pi f$, that is usually required; and if the frequency is nearly 800 c/s (actually 796) ω works out at 5,000, which is a more convenient figure than 6,283. Either is easily audible in phones, but if 50 c/s is used (as may be necessary to measure large inductances without being overmuch troubled by self-capacitance) phones are almost useless.

Detector.—For d.c. the usual detector is a sensitive centre-zero galvanometer with a variable shunt to reduce its sensitivity during the early stages of finding balance and a tapping-key type of switch to keep it out of circuit except when the key is pressed. A rectifier shunt (Sec. 5.5) is useful as a protection in case of accidents. The Sullivan portable reflecting galvanometer provides very great sensitivity at an exceptionally low cost for this type of instrument, and it is proof against damage by up to 100 times full-scale current. Nevertheless, a really sensitive galvanometer cannot be described as foolproof.

For a.c. only, phones are simple and sensitive indicators for the middle audio frequencies, though except for rough work one usually finds in the end that some amplification is necessary, especially when the surroundings are not dead quiet (Sec. 5.43). A " magic eye " without additional amplification makes a suitable detector where extreme sensitivity is not required (Sec. 5.42). In certain kinds of bridge the balance depends on frequency and so is obscured by the presence of harmonics from the source if a non-selective visual detector is used. Mental selectivity, aided by practice, comes to the rescue with audible detection, provided that the fundamental frequency is not too low. A vibration galvanometer is a very selective and sensitive detector for low frequencies, where phones fail, but is rather an expensive instrument for occasional use (Sec. 5.9).

For general purposes over a wide range of frequencies, a selective amplifier and visual indicator, such as that described in Sec. 5.23 (1), is suitable. A further refinement, facilitating balance of reactive and resistive components of impedance, is a phase-sensitive detector. In the Cossor 1446, this is used with a c.r. tube as the indicator.

Ratio Arms.—An essential part of many bridges is a pair of ratio arms; and if the bridge is to be used for d.c. as well as a.c. they must be resistances. These may conveniently be controlled by switches, of the low-contact-resistance type used for decade boxes, to give values of 1, 10, 100, 1,000, and 10,000 Ω. More important even than getting these values correct is the accuracy of ratio. And as the most accurate measurements are those made with equal ratio arms, it is especially important that corresponding resistances on each side should be equal. Fortunately it is possible to check equality by getting a balance with a pair of equal resistances—preferably of approximately the same values as the ratio arms—in the other two arms, and then reversing one pair of arms. For a.c. the ratio arms should also be as symmetrical as possible—to equalize residuals (Sec. 6.2)—and enclosed in screened boxes, as indicated in Fig. 7.2. If preferred, the resistances can be multiples of 10, one end of every one being connected to the centre terminal. This arrangement has the advantage of enabling them to be adjusted independently of one another. For methods of winding and adjustment, see Secs. 6.4 and 6.5. A typical commercial ratio-arm

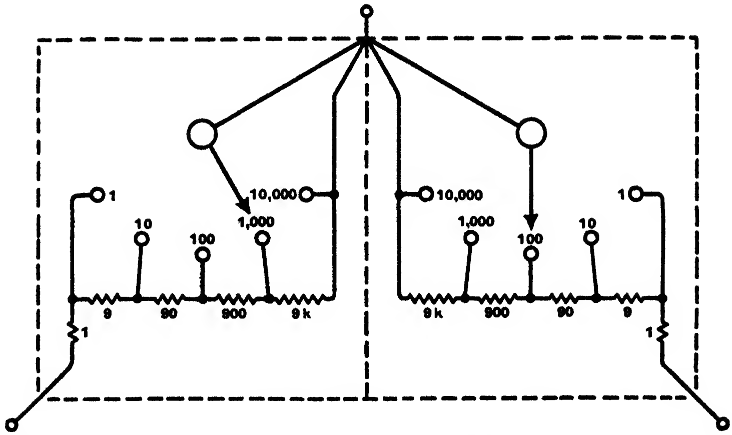


Fig. 7.2—A pair of switched ratio arms suitable as the nucleus of a general-purpose bridge

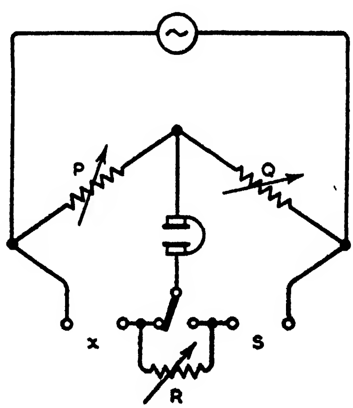


Fig. 7.3—General-purpose a.c. bridge; also suitable for d.c., if battery and galvanometer substituted for oscillator and phones. For a.c. a standard of reactance is connected at S

box is the Dawe Type 300, providing twin arms of 1, 3, 10., etc., up to 1,000 Ω.

7.4. A GENERAL-PURPOSE BRIDGE

The foregoing components—ratio arms, with source and detector for a.c. or d.c. or both—can be connected up as in Fig. 7.3 to form a general-purpose bridge that needs only the addition of standards appropriate to whatever is to be measured. In this and later diagrams the a.c. generator and phones represent any type of source and detector, not excluding battery and galvanometer where appropriate. P and Q are the ratio arms; x the terminals for the " unknown ", and S for the standard. R is a decade box, which can be used as the standard of resistance. Thus, for plain resistance measurement the switch is put over towards x and the S terminals shorted. For capacitance the standard capacitor is connected to S, and R is used to balance the resistance of the capacitor connected to the x terminals. Theoretically, the same system can be used for inductance, but in practice it seldom works very well because standard inductors in which L is variable over a wide range without altering resistance are not readily obtainable, and a bridge in which reactance and resistance adjustments affect one another wastes time and temper.

Before studying the various other types of a.c. bridge, however, one ought to consider stray admittances and residuals.

7.5. STRAY ADMITTANCES

It is easy enough to draw a diagram of an a.c. bridge, but if it shows only the intentional items of resistance and reactance it may be very misleading. The matter of residuals—the unwanted resistance and reactance that inevitably exist in every resistor, inductor and capacitor—has been examined in Chapter 6, and must not be forgotten when considering bridges as a whole; in accurate work they must be allowed for.

Another common source of error is the existence of unauthorized paths—admittances—to earth and other surroundings and between parts of the bridge itself. Assuming that the insulation is adequate, these consist of stray capacitances—but it must be remembered that they may be far from pure capacitances. Strictly speaking, there is stray capacitance from every point on the bridge to every other point on the bridge and its surroundings that is not screened off, but that is hardly a practical basis for reckoning. Since it is only the current that goes past the terminals of the arms that counts, the situation can be represented well enough by lumped capacitances between the terminals and from them to the surroundings, including earth. The problem is how to prevent them from affecting the results beyond the limits of tolerance.

There are several ways of tackling it. One method is to make sure that the undesired admittances are all very much smaller than the

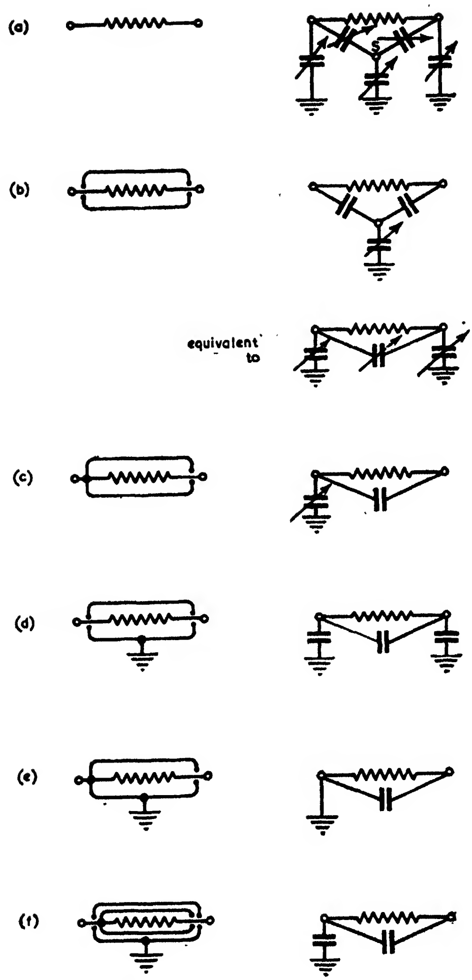


Fig. 7.4—On the left are shown various screening conditions of a bridge arm, and on the right the corresponding stray capacitances. Those whose values depend on the surroundings (S) are shown as variable

legitimate admittances of the arms. In other words, to make the impedances of the arms very much smaller than those of stray paths. If the things to be measured are inductances, one has only to use a sufficiently low frequency. E.g., at 1 kc/s a coil of 2,000 μH has a reactance of 12·6 Ω, so a shunt capacitance as large as 100 pF, which is 1·6 MΩ at the same frequency, would be quite negligible. Even 10 kc/s would be low enough for ordinary accuracy. But if the thing to be measured is a high resistance it might be necessary to use a much lower frequency to make the shunting negligible. And if it were a small capacitance there would be no advantage in lowering the frequency. But the effect of the stray capacitance might then be dodged by using a difference method (Sec. 3.9).

7.6. SCREENING

Where stray admittances cannot be ignored, they can be brought somewhat under control by screening. The object of screening is twofold: to replace indefinite and varying strays with fixed ones, and to transfer them where they will do least harm. The particular case of a capacitor was explained in Sec. 6.6. In Fig. 7.4 the same idea is presented in greater detail. The resistance symbol is meant to represent *any* kind of impedance forming an arm of a bridge. In the left-hand column are shown various conditions of screening and unscreening, and on their right the corresponding networks representing the stray capacitances. The variable ones are those affected by the position of the arm relative to the surroundings, denoted by S in the first case (*a*), which shows the situation with no screening at all. In this, all the strays vary with position, so not only are there variable capacitances from each terminal to earth but the impedance of the arm itself varies due to the indefinite capacitance across it. And all of these capacitances are liable to have large and variable loss angles.

The remaining diagrams show the effects of enclosing the arm in a fixed metal screen, which then forms the only immediate surrounding. In *b*, where the screen is " floating ", its variable capacitance to earth causes a small liability to variable shunt capacitance, as shown more clearly in the second equivalent, arrived at by the star-delta transformation (Sec. 14.14). In all the others the shunt capacitance, although inevitably present, is at least fixed, so can be incorporated in the calibration of the arm, especially if it be a capacitor. By connecting the screen to one terminal of the arm (*c*), the earth capacitance can be localized at that end. The double screen (*f*) is needed when the advantages of *e* are required without actually earthing the arm anywhere.

The principles of Fig. 7.4 can be applied to a whole bridge, and by drawing a diagram showing the strays it can be seen whether they will be unobjectionable. Fig. 7.5*b* shows the result of the partial screening of a bridge as at *a*. C_1 to C_4 represent the fixed shunt capacitances of each arm to its screen, and can be regarded as permanent parts of the arms. C_5, the capacitance between the two

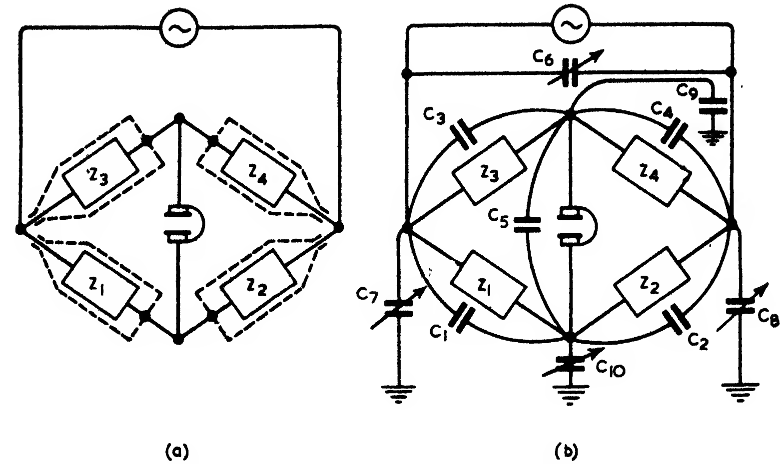


Fig. 7.5—(a) *Partial screening of a bridge, with* (b) *the corresponding stray capacitances*

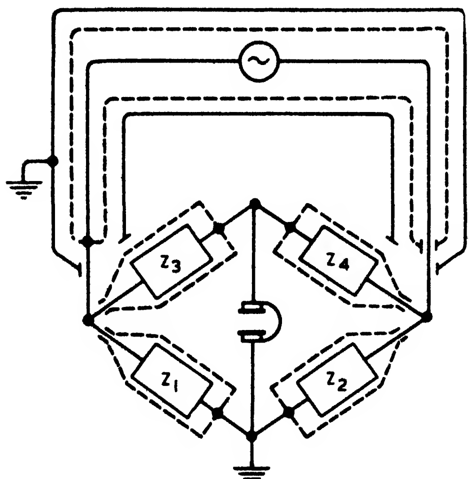


Fig. 7.6—Full screening of a bridge having one point earthed. The arm that cannot be brought to earth potential at either end (in this case the source) is screened on the Fig. 7.4f plan; the outer earthed screen is shown in unbroken line, to distinguish it from the other screens, shown dotted

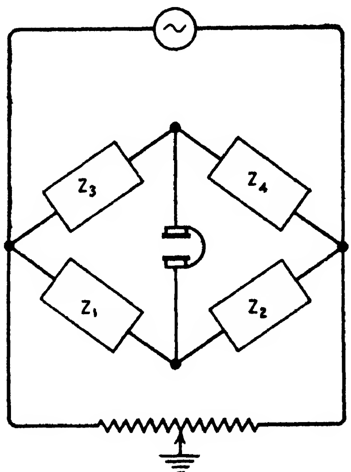
pairs of screens, does not affect the readings, because it comes across the detector; nor does C_6 across the source. But the earth capacitances from unscreened source and detector are bad. C_{10} can be eliminated by shorting it to earth, and that virtually eliminates C_9 too by bringing it across the detector; C_7 and C_8 are merged with C_1 and C_2 respectively, but they are liable to vary. If the whole source and its leads are enclosed in a double screen of the Fig. 7.4*f* type, as in Fig. 7.6, C_6, C_7 and C_8 are reduced to a single fixed shunt across either Z_1 or Z_2—in this case Z_1. In practice the whole source is not usually screened, only the secondary winding of a transformer between source and bridge.

There are several objections to this scheme, quite apart from the trouble of carrying it out. In some measurements it may not be allowable to earth and screen the item being measured. For example, one may want to measure what is called the direct impedance between two points that have other impedances (which must not be allowed to affect the reading) to earth. And if unequal-resistance ratio arms are used, and one of them is a high resistance or the frequency is high, the capacitance shunts across them may appreciably upset the ratio. And there is the interscreen capacitance across Z_1 or Z_2.

7.7. THE WAGNER EARTH

Matters can be considerably improved by bringing the detector to earth potential without actually earthing it. Suppose for the moment

Fig. 7.7—Simplest form of Wagner circuit, by which the detector can be brought to earth potential without actual earthing, thereby eliminating the bad effects of most of the stray capacitances



that a 1 : 1 ratio is being used, so that $Z_3 = Z_4$. Then at balance, when $Z_1 = Z_2$, the potential of the whole of the detector is midway between the terminals of the source, so all one has to do to bring it to earth potential is to earth the source midway between its terminals, as in Fig. 7.7. By using a sliding contact, as shown, the detector can be brought to earth potential with *any* arm ratio: a convenient method of doing so is to switch one terminal of the detector to earth and then adjust the potentiometer, or Wagner earth as it is called, until there is zero or minimum signal; then switch back and balance the bridge. In accurate work it may be necessary to repeat the adjustments several times, especially with unequal-ratio bridges, and those with reactive ratio arms. For perfect balance, in fact, the two earthing arms should have the same phase angles as Z_1 and Z_2 (or Z_3 and Z_4);

their actual impedance is preferably as low as the source can stand. Where there are large admittances to earth from the terminals of the unknown, as for example a component being measured *in situ*, a simple potential divider is inadequate.

Referring to Fig. 7.5*b*, the result of Wagner earthing is to eliminate the effects of C_9 and C_{10}, because there is no potential across them, and also C_7 and C_8, because these are merged with the Wagner arms.

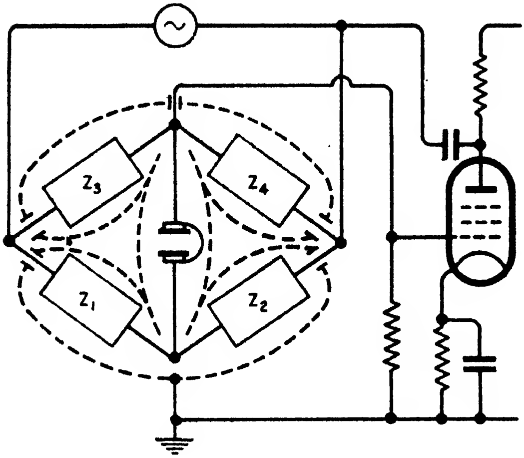


Fig. 7.8 — Automatic Wagner earth system due to C. G. Mayo

There is then often no need to do much screening, except for very-high-impedance bridge arms such as variable capacitors, and for elementary precautions to keep the detector from being directly influenced by the source.

A disadvantage of the scheme is the extra adjustment. C. G. Mayo* has described a system, used in the Muirhead Impedance Bridge, in which a valve amplifier is made to do the adjustment automatically. The effect of the capacitance between the earthed screens and the lower end of the detector in Fig. 7.8, for which 50 pF is a typical figure, is reduced approximately in the ratio of the voltage amplification, say 100, to an amount that is negligible. An alternative, applicable where resistive ratio arms are used and the stray admittances are relatively small, is to provide identical arms for the Wagner earth, the switches for the two pairs of arms being ganged.

The use of a cathode follower to reduce earth admittances is described by G. H. Rayner and R. W. Willmer in *J. Sci. Inst.*, April 1950.

7.8. SYMMETRICAL BRIDGES

The more difficult cases, such as very-high-impedance arms or high frequencies—especially r.f.—are greatly eased by using equal ratio arms and making everything symmetrical. Referring again to Fig. 7.5,

* " An Electronic Wagner Earthing Device." *Muirhead Technique*, Jan. 1947.

if the ratio arms Z_3 and Z_4, with their screens, are identical, C_3 and C_4 are equal and their effects cancel out because the phase angles of these two arms are the same. Similarly with C_1 and C_2. The arms Z_1 to Z_4, being screened, contribute nothing to C_7 and C_8, which can be equalized by connecting the source through a screened and balanced transformer; i.e., one with an earthed layer of copper sheet between the windings, the overlap being insulated to prevent it from acting as a short-circuited single-turn winding, and the two halves of the secondary wound in two halves side by side and symmetrically with reference to this screen, so that both halves have equal capacitances (C_7 and C_8) to it and hence to earth. Since C_7 and C_8 are equal, they act as a Wagner earth, bringing the whole of the detector to earth potential, even if it is not actually earthed. If it is earthed (across C_{10}), C_7 and C_8 are brought across Z_1 and Z_2, but this may be allowable provided that C_7 and C_8 are kept small, equal, and constant, especially if Z_1 and Z_2 are made the ratio arms. An earthed bridge is desirable for measuring items one terminal of which has a low impedance to earth. If the bridge is not earthed, either intentionally or through the item to be measured, it should be run as a Wagner system. C_7 and C_8 themselves, when equalized, act as Wagner arms, bringing the detector to earth potential, and their equality can be tested in the Wagner manner. But it is as well to reinforce them by lower impedances, or, even more conveniently, by using a source with an accurately balanced push-pull output, earthed at the centre, and thereby acting as its own Wagner earth. If its output impedance is low it is very tolerant of extra stray admittances and the bridge can be used to measure items, such as components *in situ*, with considerable admittances to earth from both terminals, these admittances occupying positions C_7 (or C_8) and C_{10} (or C_9). But in any cases where the admittance of C_{10} or C_9 becomes low it is advisable to make sure that the low impedance through the source is not " forcing " a balance independently of the bridge arms.

7.9. COMMON INPUT/OUTPUT BRIDGES

A major difficulty in dealing with stray admittances in any bridge based on Fig. 7.1 is that source and detector must not both be earthed. Usually the detector has considerable admittance to earth; if a pair of phones is used, it is also variable. It can be rendered harmless by earthing one of the detector terminals. If the source is a miniature battery-driven transistor oscillator, its admittance to earth can be made very small and (with the aid of simple screening) constant. But if it is a mains-driven variable-frequency signal generator, the admittance problem is much more serious.

An ingenious solution that can be useful in some circumstances has been devised by G. W. Short (*E.T.*, Dec. 1960). In what he calls a two-signal bridge, the usual detector position is occupied by a rectifier, which acts as a frequency changer. Because the detector itself is designed to respond only to frequencies different from those supplied

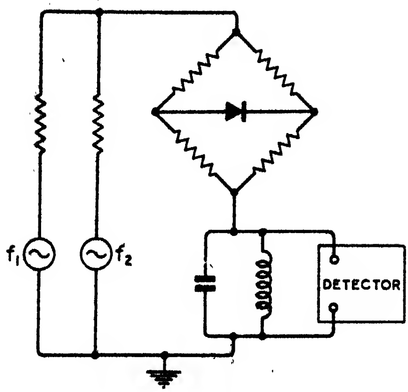
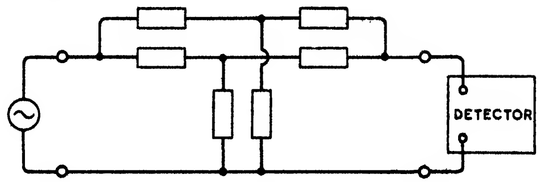
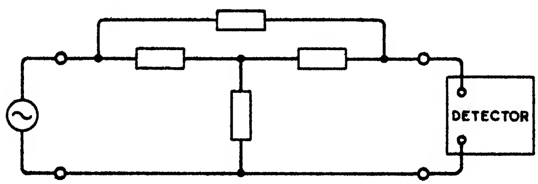


Fig. 7.9—Two-signal bridge, in which a rectifier acts as a "virtual detector", allowing the sources and actual detector to have a common earth terminal



(a)



(b)

Fig. 7.10—The parallel-T (a) *and bridged-T* (b) *networks also allow both source and detector to be earthed, and are similar in many respects to bridges*

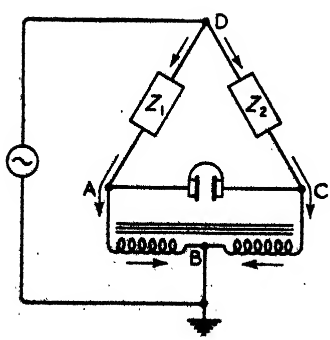


Fig. 7.11—Principle of bridge based on transformer ratio arms, which give the benefits of Wagner earthing without any adjustment

(f_1 and f_2 in Fig. 7.9) it can be connected in series with the source as shown, with one of its terminals common and earthable. The sum and difference frequencies are generated only when f_1 and f_2 currents pass through the rectifier; i.e., when the bridge is unbalanced to them both. The scheme can be operated with a single source, the sum and difference then being $2f$ and zero, but that obviously limits the possibilities.

In Fig. 7.9 the rectifier acts as a " virtual detector ", but in a modified form of circuit it is a " virtual oscillator ". In another circuit variation, input and output signals can be carried by a single line—a facility that offers possibilities in telemetry. Details of these and other arrangements are given in the article cited.

Although not strictly bridges, the parallel-T and bridged-T (Fig. 7.10) are null measuring networks with a wide range of usefulness combined with the advantage that source and detector have a common terminal, which can be earthed. The principles were explained by W. N. Tuttle (*Proc. I.R.E.*, Jan. 1940) and their application in a General Radio instrument was detailed by D. B. Sinclair (*Proc. I.R.E.*, July 1940). More recently the advantages of the bridged-T network were considered by J. K. Choudhury and P. C. Sen (*E. & R.E.*, Nov. 1959) with particular reference to magnetic measurements. Besides the common earth point, the advantages are that no multiple screening is needed, even at radio frequencies, and at balance the entire supply voltage appears across the impedance being measured, so that measurements can be made on iron-cored coils with a sinusoidal flux waveform.

7.10. TRANSFORMER RATIO ARMS

Finally, the following method, due to A. D. Blumlein, of coping with stray admittances is in some ways the best, and has enabled hitherto impossible measurements to be made—such as capacitances as low as 0·000001 pF at a.f.,* without disconnecting them from relatively low impedances to earth; and bridge measurements at 200 Mc/s.

Fig. 7.11 shows the delightfully simple principle. ABC is a coil in which the coupling between sections AB and BC is as near 100% as possible. In Blumlein's original design they were bifilar windings; that is to say, the two wires were wound on together like one. Then when Z_1 and Z_2 are in the same ratio as the turns in AB and BC (i.e., at balance) the core is subjected to equal ampere-turns of opposite sign, so is unmagnetized and no voltage is generated across any part of the coil. In fact, the only voltage across the coil is due to the currents flowing through its resistance and leakage inductance, and these are deliberately made very small. Virtually, then, A and C as well as B are at earth potential, and admittances to earth from those ends of Z_1 and Z_2 and from both ends of the detector have negligible effect, even when quite large. Admittances from their D ends merely shunt

* " A Direct Capacitance Aircraft Altimeter ", by W. L. Watton and M. E. Pemberton. *Proc. I.E.E.*, Part III, May 1949.

the source. Such a bridge can therefore be used for measuring quite small capacitances without disconnecting the component from its circuit, if that is earthed. Besides that great advantage, ratio-arm residuals are absent, and the ratio is determined exactly by the number of turns, so is more quickly and precisely obtained than by winding and adjusting resistive arms. And it is obviously constant. On the other hand, the coil has to be rather specially made if it is to approximate the ideal sufficiently. The subject has been treated by F. Butler

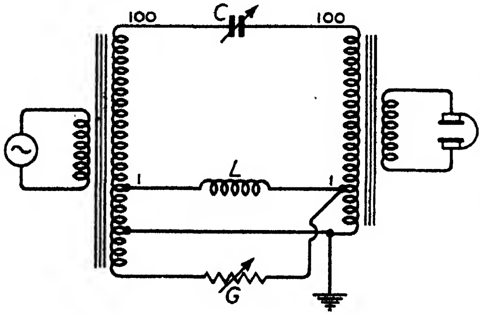


Fig. 7.12—Showing how a second transformer can be used in the detector arm to increase the overall ratio—in this example, 10,000 : 1. Inductance (L) is measured on the same side as the standard capacitance C, its resistance being balanced by the conductance G

(*E.T.*, Aug. 1960). By taking great care, the N.P.L. claims to have constructed transformers with ratios correct to 1 in 10^7.

As with other bridges, the source and detector can be interchanged. An advantage of doing so is that the ratio arms and source-impedance-matching transformer can be combined, by putting an extra winding on the ratio-arm core. Since points A and C are then clearly not " earthy ", the beneficial principle as regards admittances from them to earth is less obvious. But an admittance across any part of a perfect transformer loads the whole, so the ratio is unaffected. In practice, with careful design, ratios as high as 1,000 : 1 can be provided with good accuracy.* The design of a bridge for low capacitances ($<$1,111 pF) has been dealt with by J. F. Golding and L. G. White (*M.I.*, June 1960).

A number of tapping-points can be provided on the transformer, to give various multiplying ratios. And L can be measured against C, reckoned as negative L, by tapping both on the *same* side of B. To extend the range of measurement beyond the ratio that is practicable with a single transformer, the detector also can be connected through a multi-ratio transformer. Fig. 7.12 shows an example, in which an unknown L is balanced by a standard variable C; and as both source and detector are tapped in the ratio 100 : 1 the overall ratio is 10,000 : 1. The detector can be tapped across any proportion of the transformer that suits its impedance. This kind of bridge is really an admittance bridge, so balance in this example means that the admittance of L is $-10,000$ times that of C. In order to find L it is necessary to know

* " Double-Ratio A.C. Bridges with Inductively-Coupled Ratio Arms ", by H. A. M. Clark and P. B. Vanderlyn. *Proc. I.E.E.*, Part III, May 1949.

the frequency, and for convenience this should be chosen so that ω, $= 2\pi f$, is a round number such as 10,000 ($f = 1{,}592$). Then $1/10{,}000\ L = 10{,}000 \times 10{,}000\ C$, so $L = 1/10^{12}\ C$, or $1/C$ if C is in pF. If C were 200 pF, L would be 0·005 H. Since C and especially L would not be pure reactance, their phases would be less than 180° apart, and perfect balance would be impossible without some conductance G on the other side of the bridge. This is considered as conductance rather than resistance, because it represents the in-phase path *in parallel* with L. The higher the Q of the coil, the higher the resistance.

Although so simple in principle, this kind of bridge calls for a good deal of thought in its details; for example, in the design of the transformers, and the means for balancing conductive and susceptive parts of admittance over wide ranges. Principles and practical details have been discussed by J. F. Golding (*W.W.*, June 1961). The bridge can be earthed or not, according to requirements; generally it will need screens, connected to point B.

7.11. RESISTANCE AND CAPACITANCE BRIDGES

Having studied the stray admittance problem one can more intelligently consider the diverse merits of what a gardener would call the named varieties of bridges. The most important are laid out for comparison in Fig. 7.13, with their equations for balance. There is no need to show source and detector; these are connected across the diagonals, and it is optional which goes where. Except in *a*, the unknown is represented by its equivalent series resistance r_x and capacitance C_x or inductance L_x, enclosed by a dotted line to show that they are inseparable. They can always be converted to parallel equivalents if desired (Sec. 14.12).

The simple bridge described in Sec. 7.4 (Fig. 7.3) can be used either as *a*, which requires no comment, or *b*, which is a very useful type. If there is no need to provide for d.c. as well as a.c., it could be improved upon by substituting inductive ratio arms. Without much difficulty it can be made to read to a fraction of 1 pF. For capacitances up to the limits of a variable air capacitor it is helpful to use equal ratio arms, Wagner earthing, and substitution or difference method of measurement; and of course a screened standard. For larger capacitances these precautions are less necessary, and R is in general smaller.

The same bridge can be used for comparing inductances, but in practice seldom is, for the reasons given in Sec. 7.4.

When C_x is very large, r_x may be very small and it may be more convenient to balance with a comparatively large resistance in parallel, as in *c*. The equations are more complicated because they incorporate the series-parallel transformation, but if the unknown is reckoned in terms of its parallel equivalents the equations are the same as for *b*. If parallel R and C are balanced against an actual pair in series, balance is (as the equations show) obtainable at only one frequency, and the

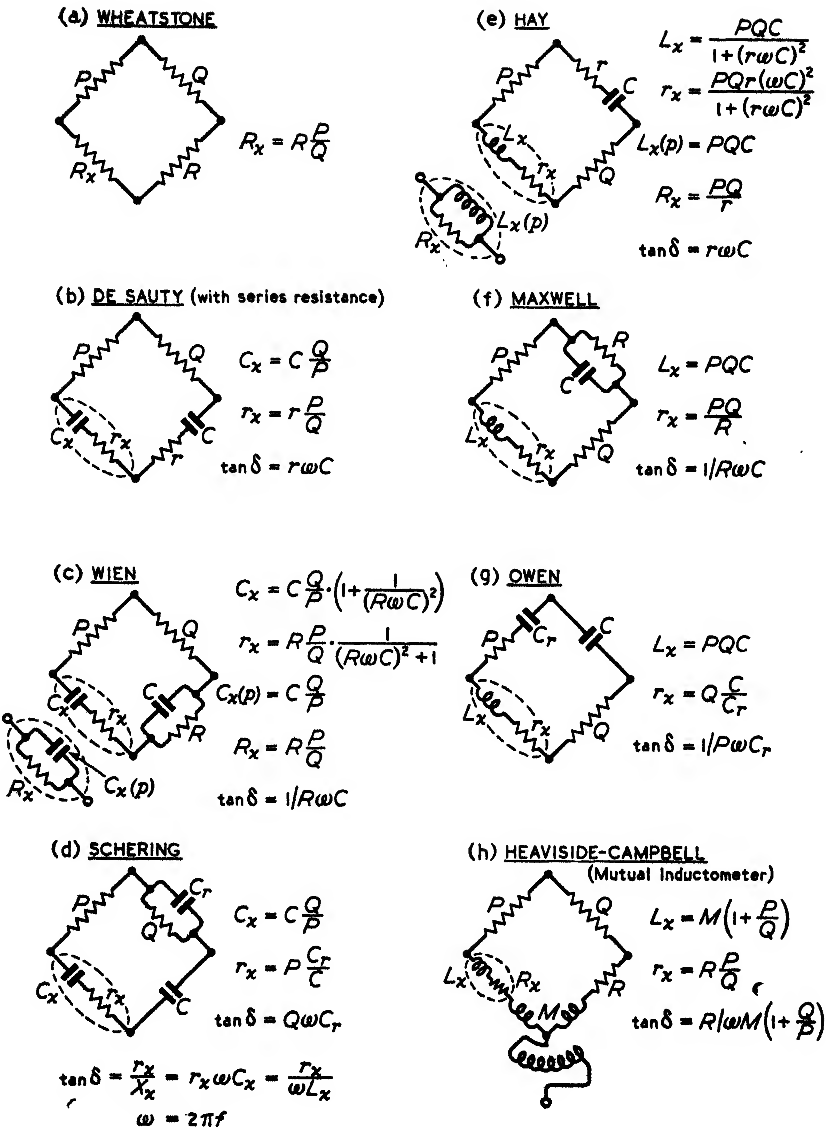


Fig. 7.13—Summary of the most-used named varieties of bridges, with their balance equations. Source and detector are omitted for clearness

Wien bridge is more often used with known capacitance and resistance to measure frequency than vice versa (Sec. 7.19).

The most generally favoured bridge for accurate measurements of capacitance and phase angle, and properties of dielectrics, especially at very high voltage, is the Schering (*d*). The difference between it and *b* or *c* is the method of balancing r_x, by a variable capacitor across the opposite arm Q instead of by resistance in series or parallel with the adjacent standard, C. The advantage is that, if necessary, all adjustments can be carried out on the P and Q arms, earthed at their junction, so that if the junction of the other two is connected to a very high voltage nearly all this voltage comes across those arms and there is no danger to the operator.

7.12. INDUCTANCE BRIDGES

Reactances of opposite kinds can be balanced against one another if they are in opposite arms. So a standard of capacitance (which in general is better than a standard of inductance) can be used for measuring both capacitance and inductance. All that is needed to convert type *b* or *c* into an inductance bridge is to transpose the Q and standard arms, giving types *e* and *f* respectively. But note that it is now the parallel-standards type (*f*) that has the simpler equations. Type *e* is better for measuring large inductances, particularly iron-cored; and for the parallel equivalents of the unknown the balance equations are the same as Maxwell's.

The Maxwell equations show that ideally L_x ought to be balanced by adjusting C, and r_x by R. It is not practicable to vary C continuously much above 1,000 pF, which at a.f. is a very high impedance (and therefore vulnerable to strays) and necessitates extreme ratios for measuring low L_x (or large C_x). Also it leads to awkwardly high values of R. On the other hand, adjusting P or Q affects both L_x and r_x balances at once, which is apt to be frustrating. These disadvantages can be mitigated by thoughtful design.

The Hay bridge is particularly suitable for measuring the inductance of iron-cored coils with d.c. flowing, because the d.c. source can be connected in series with the coil.

The Owen bridge (*g*) differs from Maxwell in balancing r_x by capacitance in series with P instead of resistance in parallel with C. Ideally, C_r and P are the balancing controls. C can conveniently be quite a large fixed capacitance, and Q varied in steps for range changing. P can be varied continuously for L_x balance, and the difficulties of varying C_r can be reduced by loading up the x arm with extra resistance.

Possibly the best bridges for inductance at frequencies not above about 1 kc/s are those based on mutual inductance as the standard. Mutual inductance has the advantage over self-inductance that it is pure reactance, provided that the frequency is low enough for self-capacitances not to introduce an appreciable in-phase component; and it can be varied down to zero and beyond, into negative values. So it can be used for measuring capacitance too. The arms have such low impedance that stray admittances can usually be ignored, but reasonable care is needed to see that the various coils are placed so that they do not couple inductively. This difficulty is sometimes reduced to a minimum by the use of toroidal windings. The only serious drawback is the high cost of mutual inductometers.

The Campbell standard mutual inductometer, made by the Cambridge Instrument Co., shown in Fig. 6.14 and described in Sec. 6.10, is a high-precision instrument and is provided with facilities enabling it to be used for measuring mutual inductance, self-inductance from fractions of 1 μH to thousands of henries, with superimposed d.c. if necessary, and capacitance over a wide range. In the Heaviside-Campbell bridge (Fig. 7.13*h*) the ratio resistances P and Q are equal, and this is preferred for high accuracy. The inductometer fixed coils are connected in the other two arms, and having equal self-inductance they balance. L_x is balanced by mutual inductance M, injected by adjusting the moving coil, which is in series with the source. Up to rather more than an ohm, the r_x balancer R is provided in the inductometer by a slide wire at the junction of the coils, so additional R, where shown, need not be varied in smaller steps. Coils are provided for ratios of 10 : 1 and 100 : 1, and to suit these P and Q should include the values 10, 90, and 990 Ω. With unequal ratios the slide wire is out of action and the external R must be continuously variable.

7.13. COMMERCIAL BRIDGES

For ordinary component checking over a wide range of C and R, a basically De Sauty bridge in which balance is obtained by continuously variable ratio arms consisting simply of an ordinary " potentiometer ", with 50-c/s mains as source and " magic eye " as detector, is obtainable in a number of different makes (e.g., Hunts) and also as a Heathkit (Model C-3U). Fixed standards of C and R are selected by a range switch, and a position is often provided for comparison with an external standard. At such a low frequency there is little trouble with stray admittances, and little or no screening is needed. And except where substantial reactance is present, one can measure what is virtually the d.c. value of a resistor. The Heathkit costs less than £8, has two ranges of resistance covering 100 Ω to 5 MΩ, and four of capacitance from 10 pF to 1,000 μF, and measures leakage and power factor of electrolytic capacitors at five polarizing voltages from 5 to 450. Constructional details of another instrument in this category are given in the next Section.

Most of the leading makers also offer a " universal " bridge, measuring a wide range of L, C and d.c. R with an accuracy of the order of 1%, and also tan δ or Q. Typical examples are the Marconi Instruments TF868B, the General Radio 1650A, the Cossor 1446 and the Avo Type 1. All of these use the ordinary Wheatstone network for d.c. resistance, but the d.c. detectors differ. Only the G.R. bridge uses the traditional unaided pointer instrument. The Avo uses a d.c. amplifier, and the two other makes get round the problem by converting the out-of-balance d.c. to a.c. with a chopper, thereby utilizing the a.c. amplifier-detector. All four have visual indicators instead of (or in addition to) phones. G.R. and Avo a.c. sources are 1 kc/s; Cossor, 2 kc/s; M.I., 1 kc/s and 10 kc/s. For capacitance, Cossor use De Sauty; Avo, Wien; and the other two use both, according to whether Q or tan δ are measured. All four use the Maxwell circuit for inductance, but the M.I. and G.R. provide also for the Hay. Because versatile bridges are almost bound to make the main measuring control a variable resistance, one looks with interest to see how the chief snag of the Maxwell bridge—interdependent inductance and resistance balances—is avoided. M.I. and Avo measure Q or tan δ rather than r_X. As the formulae in Fig. 7.13 show, this does not involve R_2 or R_3, either of which can accordingly be used for independent determination of L. The Cossor indicator is a small c.r. tube, the X deflection of which is derived straight from the source and the Y deflection from the bridge. It is therefore phase-sensitive as well as amplitude-sensitive, facilitating adjustment of the phase balance (uncalibrated). The G.R. bridge is calibrated in Q, and for enabling balance to be reached quickly even with fractional-Q components it incorporates the "Orthonull" device, whereby adjustment of R_2 carries R along with it, but this ganging does not work in reverse. A full explanation is given in *G.R.E.*, April 1959.

Marconi Instruments also offer a similar bridge with higher precision ($\frac{1}{4}$%) and wider ranges—the TF1313.

The Cambridge universal bridge, not surprisingly, uses a decade mutual inductometer, which is of toroidal construction, and is accurate to within $\pm$ 0·1 %.

For the measurement of inductance only there is a lot to be said for the Owen network, and this was the General Radio choice in a recent introduction, Type 1632-A, described in *G.R.E.*, Nov. 1959. The accuracy over most of the wide range is given as 0·1%.

Wayne Kerr have made good use of the advantages of the Blumlein bridge. Their Type B.521 already mentioned has Fig. 7.12 as its basis, and uses the 50 c/s mains as source and a selective amplifier and "magic eye" as detector. The accuracy is 2% for general *in situ* component measurement. Type B.221, with a 10 kiloradians/sec source, gives higher accuracy—$\frac{1}{4}$%. Type B.321, 10 kc/s, with similar accuracy, is specialized for low inductance and resistance, 0·002 μH to 0·1 H and 0·02 mΩ to 1 kΩ. It uses a modified circuit, similar to Fig. 7.12 on the detector side, but with the source connected across the unknown in series with a standard resistance, the voltages across these being balanced against one another respectively through a fixed resistance and decade resistances and capacitances in parallel. Lastly, wide-band iron-cored 1 : 1 r.f. transformers used in Type B.801 enable admittances to be measured at frequencies from 1 to 100 Mc/s. The source and detector are separate units connected by

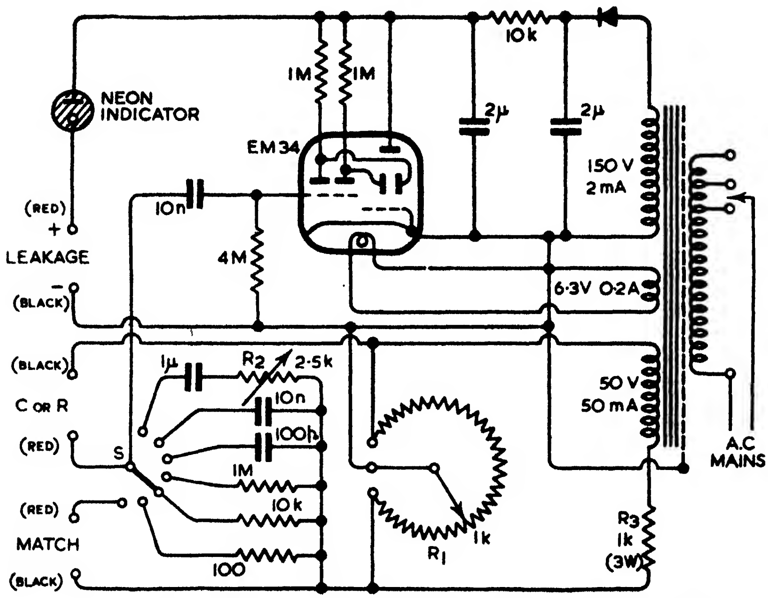


Fig. 7.14—Circuit diagram of mains-driven bridge for capacitance and resistance, using visual balance indicator. The rectifier is a Westinghouse type 16HT12

coaxial cables. A monograph on the principles of the transformer ratio-arm bridge, by R. Calvert, is obtainable from this firm.

Cintel have two transistorized transformer-ratio bridges; one for capacitance over a wide range, the other for self and mutual inductance.

A good example of the use of the Hay network for incremental inductance (with d.c. flowing) is the one by Furzehill.

7.14. CONSTRUCTION OF A MAINS-FREQUENCY BRIDGE

One advantage of bridge operation at 50 c/s is that a very adequate oscillator is maintained by the electricity supply authority. Another is that the influence of stray capacitances is much less than at 1 kc/s or more, and for most purposes the precautions discussed in Secs. 7.5 to 7.8 are almost or entirely unnecessary. But unfortunately the reactance of r.f. coils is too low to be easily measurable. If low-inductance measurements are abandoned, a very useful method is to use an ordinary " potentiometer " as variable ratio arms for comparing the unknown with one of a number of fixed standards. Fig. 7.14 shows a circuit suitable for amateur construction, and Fig. 7.15 a complete instrument. The scale of the potentiometer, R_1 in Fig. 7.14, is calibrated to read ratios directly, as in Fig. 7.16. The scale reading at balance, multiplied by the value of the standard selected by the switch S, gives the value of the component connected to the " *C* or *R* "

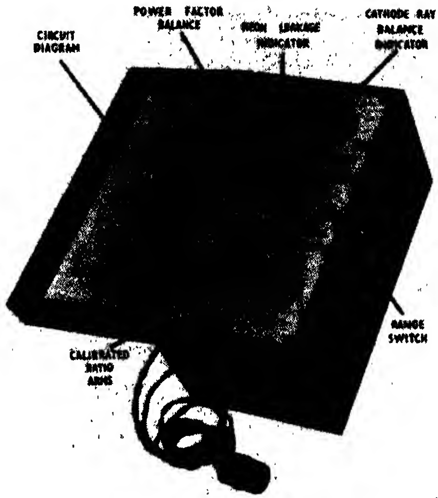


Fig. 7.15 (right)—Example of embodiment of bridge circuit shown in Fig. 7.14

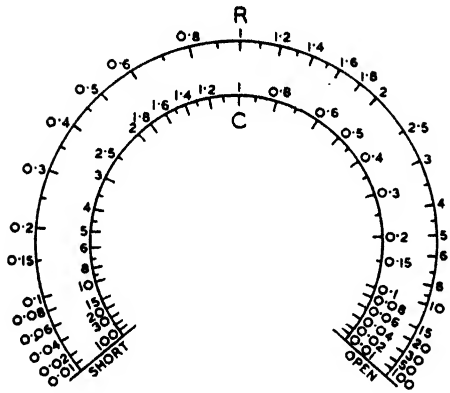


Fig. 7.16 (left)—Scale of the type of bridge in Fig. 7.14, assuming a linear R_1

terminals; and as in this instrument all the standards are multiples of 10 the mental arithmetic demanded is not severe. For example, if the reading when balance is obtained is 0·65, and the range switch is set to 10 k, the answer is 6·5 kΩ. Balance is shown by a cathode-ray tuning indicator (magic eye), and the EM34 has one sensitive shadow for use near balance and another, less sensitive, for initial adjustment.

When made as specified, the instrument measures resistance from 10 Ω to 10 MΩ, and capacitance from 10 pF to 10 μF, including electrolytics; and gives rough checks over considerably wider ranges. It measures power factors up to 60% of capacitors from 0·1 μF upwards. It compares components, including large inductances, with any standard of similar sort, detects leakages from about 100 MΩ downwards, and gives a continuously variable 50-c/s signal up to 25 V.

It is portable, self-contained (except for mains connection), direct-reading, and requires no earth.

The signal voltage applied to the component under test should be graduated to suit its impedance, and this is done automatically by R_3. For high impedances such as grid leaks and small capacitors, which would be difficult to balance sharply without adequate signal voltage, 25 V is available; but for impedances through which this voltage would pass too much current for themselves and the transformer, the voltage falls to a suitable value. R_3 has been chosen so that even if the test terminals are shorted no harm will follow. And electrolytic capacitors rated at low voltages will not be damaged.

R_1 should be a reliable component with a smooth and uniform linear resistance element, and preferably larger in diameter than the ordinary volume-control type. Since it is only the ratio that is calibrated, the exact total resistance is not important.

The accuracy of measurement is of course no better than that of the standards used. A suitable tolerance for them in this instrument is $\pm$ 1%. The 100 Ω should be wire-wound; the 10 kΩ may well be, if non-inductive, but the 1 MΩ would be a high-stability carbon resistor. A ceramic capacitor is suitable for the 100 pF, and a good-quality mica for 10 nF; but the cost of mica is prohibitive for 1 μF so the best obtainable paper may be used. If not " spot-on ", the capacitors selected should be slightly low, so that they can be brought up to the exact value by small ones across them. With the standards specified, the scale is used from 0·1 to 10. If balance falls outside these limits a better result will be obtained on another range. Near the limits, the readings are rather cramped, and the most worth-while improvement one can make is to add standards of 1 kΩ, 100 kΩ, 1 nF, and 0·1 μF. With this augmented selection, the standards can be checked against one another.

No provision is made for balancing the resistance of the smaller capacitances. It is generally enough to notice whether the sample under test gives a reasonably sharp balance; if not, it is a bad one. But most of the largest sizes are electrolytic, and a power factor of, say, 5% is not necessarily a ground for rejection. R_2 is used to measure power factor on the 1-μF range. A sharp balance should be sought, by alternate adjustment of R_1 and R_2. Strictly, a polarizing voltage should be applied to electrolytic capacitors when they are being measured, but readings taken without agree well with those obtained under more orthodox conditions.

When using the 100-pF range the same sharpness cannot be expected as on the others. It is not that balance is blurred; assuming the power factor is satisfactory, balance is quite clear, but a larger movement of R_1 is needed to show appreciable change in the indicator illumination. The best way is to swing the knob to and fro between positions each side of balance until the exact centre point can be judged. If care is taken with the layout on this range, and possible displacement of zero allowed for, a capacitance as low as 5 pF can be observed, and readings around 100 pF are surprisingly good. The 100-pF and 1-MΩ standards should be joined close to S and have minimum capacitance to other parts. All a.c. leads must be kept as far as possible from the grid of the valve.

A useful feature is a spare position on the range switch, for connecting an external standard to the " Match " terminals. For example, it might be necessary to adjust a number of components to match a pattern. This facility is usable for inductances, if not less than about

Table 7.1

Scale reading (per cent)	R_2 (*ohms*)	
	Power factor	*Dissipation factor*
5	160	160
10	320	320
15	485	480
20	650	640
25	820	800
30	1,000	950
35	1,190	1,110
40	1,400	1,270
45	1,610	1,430
50	1,830	1,590
55	2,080	1,750
60	2,370	1,910

1 H. Push-pull transformer windings can be tested for equality. And other uses will readily occur.

For calibrating R_1 it is a great help if one can use a laboratory resistance box and at least one accurately-known resistance. These are connected to " *C* or *R* " and " Match ", and their ratios set to various values. The *C* scale, which is the reciprocal of the *R* scale, can be obtained by repeating the process with the two resistances interchanged. Alternatively, the same scale can be used, if provision is made on the range switch to reverse the potentiometer on *C* ranges. If no resistance box is available, a variable resistance can be equalized to a fixed one by adjusting until the same setting (to be calibrated " 1 ") of R_1 gives balance when the two are interchanged. If a number of variables are available, this process can be repeated and the equal resistors thus obtained connected in series to give points 2, 3, etc.

To calibrate R_2 in power factor, the 1 μF is temporarily shorted and R_2 is balanced against known resistance in the " *C* or *R* " arm. Table 7.1 gives the required values of R_2 in series with 1 μF at 50 c/s for a power-factor scale, and also for a dissipation factor scale (Sec. 6.2), in case it is preferred.

The leakage test is an optional extra, requiring only a neon lamp. A suitable type of lamp is that supplied for indicator purposes. The lowest voltage rating should be chosen, but it may be noted that a 230-V lamp lights up on the voltage available. Though it can be used for detecting d.c. leakage anywhere, its main use is for capacitors. When one is connected, the lamp flashes momentarily due to charging, and in a dim light this is perceptible with capacitances down to 5 nF or even lower. If the capacitor is a good one it may be necessary to wait quite a long time for the next flash. But a poor one flashes like a police beacon, while a downright bad one shows a continuous light. This judgment must not be applied too literally to electrolytic capacitors (which must always be connected the right way round), because they always leak to some extent, but whether or not that extent is reasonable can be gauged by watching the lamp. A light that remains bright after an electrolytic capacitor has been connected for a minute or so shows excessive leakage. Frequency of flashing depends not on megohms alone, but on megohm-microfarads, a quantity that is a truer measure of merit; 50 MΩ leakage in 1 μF is equivalent to 200 MΩ in 0·25 μF.

The variable 50-c/s signal mentioned is obtainable between " Leakage— " and one of either of the other pairs. And of course a 25-V signal with a variable tapping is available by using all three terminals.

Crocodile clips are quicker to use and generally more convenient than screw terminals; the advantages of both are obtainable by having the clip leads attached to terminals. Matt ivorine sheet is suitable for covering the panel, as it looks good and takes pencil calibration which can afterwards be made permanent with Indian ink. But remember it is inflammable!

7.15. ADAPTOR FOR IRON-CORED INDUCTORS

With a few additional components the *C*-and-*R* bridge just described can be adapted to deal with inductance from about 1 to 1,000 H, provided that the power factor is small. Air-cored coils are therefore practically ruled out. Measurement can be made with d.c. flowing, up to 100 mA if not kept on too long. The scale can be calibrated from the main bridge. The addition can either be incorporated, with a switch for disconnection, or connected externally as an adaptor. Assuming the latter, Fig. 7.17 shows the circuit and mode of connection.

The transformer is a Bulgin LF.45 or other miniature 1 : 25 ratio, for increasing sensitivity of indication. C_3 is to reduce harmonics that would blur the balance. The 0·25-μF capacitor C_1 is the standard, so ought to be of reasonably good quality. The total resistance R_4 is not critical (nor is R_5) but it must not vary more than 1 Ω as the slider is moved. A Morganite carbon potentiometer has been found the best solution. C_2 is to prevent current from the polarizing battery from passing through the transformer.

R_4 is first set to short-circuit C_1 and the main bridge is then balanced (at approximately 1 on its scale). The coil to be measured is then connected to the terminals "*L*," in series with a battery and milliammeter

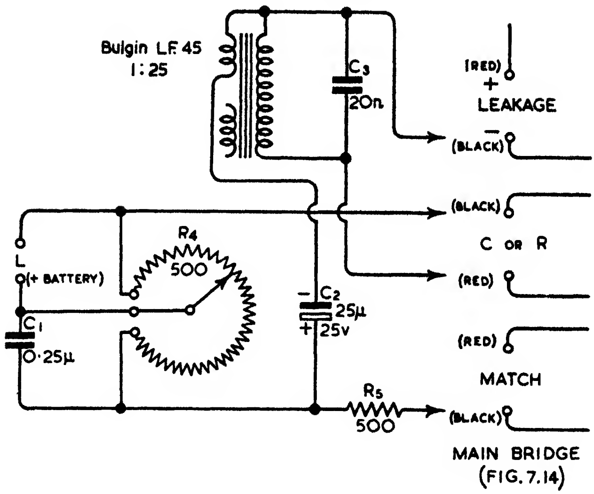


Fig. 7.17—Adaptor for measuring iron-cored inductors, showing how it is connected to the bridge shown in Fig. 7.14

if polarizing d.c. is required. (The + terminal of the battery must go to the L terminal so marked.) R_4 is then adjusted to rebalance. To obtain it sharply it may be necessary to adjust the main bridge potentiometer, but if such adjustment exceeds 2 or 3% it is a sign that the resistance of the coil is higher than normal for iron-core types, and the reading cannot be relied upon.

At balance the inductance L in henries is $\dfrac{(a/b)^2}{4\pi^2 f^2 C_1}$ where f is 50 c/s, C_1 is in this case 0·25 μF, and a/b is the ratio of the resistance across L to that across C_1 and can be calibrated from the main bridge, which, of course, is itself scaled in ratios. The formula simplifies to $L = 40\,(a/b)^2$,

Table 7.2

Ratio on main bridge R scale	0·16	0·22	0·32	0·50	0·71	1·00	1·58	2·74	3·16	5·00
Corresponding calibration in henries	1	2	4	10	20	40	100	200	400	1,000

so the 500-Ω potentiometer can be calibrated in henries by connecting its slider to the " red " terminal (grid of EM34) on the main bridge and its ends to the " black " terminals (ends of main potentiometer) and applying Table 7.2.

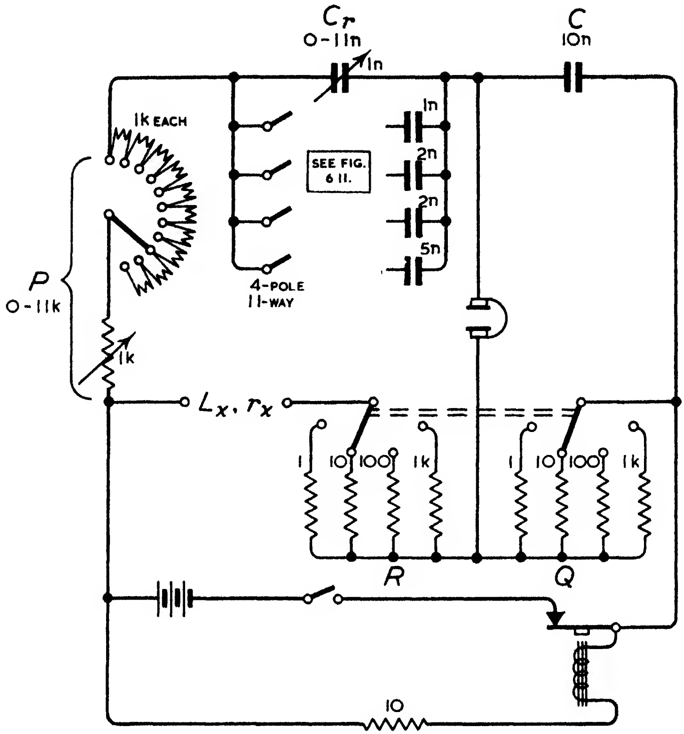


Fig. 7.18—Owen bridge for inductance, incorporating improvements by L. B. Turner. The lettering corresponds to Fig. 7.13g

7.16. CONSTRUCTION OF AN INDUCTANCE BRIDGE

The two preceding sections cover most values of resistance and capacitance, and the iron-core or a.f. range of inductance. There only remain inductances below 1 H. The 50-c/s bridge fails here, chiefly because the reactance of a few μH at that frequency is so minute as to be negligible in comparison with its resistance. The inductometer bridge is good but expensive. So the Owen, in which the standard is a fixed capacitor, deserves consideration.

An adaptation of the Owen bridge, due to L. B. Turner, is easily and inexpensively made. The circuit (which should be compared with Fig. 7.13*g*) is shown in Fig. 7.18. *R* and *Q* are varied simultaneously by what is in effect a range switch. If the *L* terminals are shorted, then for balance the variable C_r must be made equal to the fixed *C*, and *P* set to zero. Connecting inductance between the *L* terminals makes it necessary to introduce a proportionate value of

P; and to balance the resistance (r) of the coil the reactance of C_r must be increased, and therefore its capacitance reduced.

C is the standard and must have a low loss, but its opposite number, C_r, need not be particularly good and may consist of an ordinary variable of at least 1 nF, supplemented by switched capacitors (which may be ordinary receiver type) to total 10 nF. Fig. 6.11 shows how a four-pole 11-way wafer switch can be used to obtain this in 1-nF steps using only 4 capacitors. P is also not necessarily very pure; ordinary winding with periodical reversal will do. Reactive residuals merely affect the setting of C_r slightly. R and Q, however, ought to be non-inductively wound, as described in Sec. 6.4.

Since the balance condition is $L = PQC$, the maximum values in the four ranges are 0·1, 1, 10, and 100 mH; so the total range covered is from a few μH to 0·1 H. The arrangement is not so well adapted for measuring r; it is $QC/C_r - R$, or (since $R = Q$), $Q(C/C_r - 1)$. To provide accurate measurement, P would have to be non-reactive and C_r known.

A special feature is the method of connecting the battery, enabling a current of the order of 1 A to be passed through L_x on the lowest range. A power transistor oscillator would be even better. The impedance of the bridge being low, that of the phones should be likewise.

The author has made a successful bridge on similar lines but using a valve oscillator of about 2,000 c/s, with range arms 5, 50, and 500 Ω, and P consisting of a good 500-Ω "potentiometer" and 10 steps of 500 Ω each. It gives the useful ranges of 0·25, 2·5, and 25 mH, and a coil of 1 μH is measurable.

Constructional details of a Hay bridge for measuring iron-cored coils, if necessary with a known amount of d.c. flowing, are given by F. E. Terman in *Measurements in Radio Engineering*, pp. 53–57, and by Terman and Pettit in *Electronic Measurements*, 2nd edn. (McGraw-Hill), p. 111.

7.17. R.F. BRIDGES

There is, of course, no difference in principle between r.f. bridges and a.f. bridges, but the higher frequency accentuates the difficulties discussed in Secs. 7.5 to 7.8, and also calls for appropriate source and detector, which can conveniently be a r.f. signal generator and a receiver respectively. Thorough screening of the bridge and its connections is essential and must be carefully thought out. Transformer ratio arms have been found particularly valuable; the Wayne Kerr v.h.f. bridge of this type (B.801) has been mentioned in Sec. 7.13. The techniques are outside the scope of this book, and the following selection of references may be helpful:

Radio Frequency Measurements, by L. Hartshorn (Chapman & Hall).
Radio Engineering, Vol. 2, Ch. XX, by E. K. Sandeman (Chapman & Hall).
"A New Technique in Bridge Measurements", by R. Calvert. *E.E.*, Jan. 1948.
"Experimenters' R.F. Bridge", by H. V. Sims. *W.W.*, May 1952.

7.18. INFLUENCE OF SOURCE FREQUENCY

Looking at the bridge balance conditions set out in Fig. 7.13 we see that most of them contain no reference to frequency. Provided that the bridge arms are as shown, in fact as well as in theory, and that the values of their components are unaffected by frequency, the frequency of the source makes no difference to the balance. This is convenient, because it means that there is no need for the frequency of the source to be known, nor need its waveform be perfectly pure. In practice, however, the values of components, especially the resistance of coils, may vary somewhat with frequency, and arms that seem to have the same form in the diagrams may differ—e.g., coils are complicated by self-capacitance—so that they may not be exactly equivalent at more than one frequency. When the bridge is balanced for the fundamental frequency, therefore, it is more or less out of balance for the harmonics. Unless the source waveform is pure, or the detector is able to reject the unwanted frequencies, it may be difficult to find exact balance. An untuned visual detector is upset by even a small amount of these. With practice the ear can discriminate between fundamental and harmonics, but is never so sensitive in such conditions.

In bridges where the arms compared have different frequency characteristics, such as the Wien (Fig. 7.13*c*) and Hay (*e*), the frequency must be accurately known and the source waveform very pure or the detector sharply tuned. This difficulty can be exploited for exploring the nature of the item under test, instead of finding its simple series or parallel equivalent, effective at only one frequency. The method, described by Thomas Roddam in " New Bridge Technique " (*W.W.*, Jan. 1950), consists in using a square-wave signal—comprising a wide range of frequency—and adjusting not only the values of the standard arm but also its network form until balance, indicated by a c.r. oscilloscope, is complete. He made use of the transformer type of bridge, as did K. Lamont (*E.E.*, Aug. 1955) in developing this method for the measurement of iron-cored inductors.

Other uses made of frequency-discriminating bridges include frequency measurement, and the filtering out of one particular frequency, as in the measurement of distortion.

7.19 FREQUENCY BRIDGES AND METERS

The most-used frequency bridge is the Wien (Fig. 7.13*c*). The signal whose frequency is to be measured is applied as the source, and " C_x " and " r_x " are incorporated in the bridge. If they are varied simultaneously with C and R to obtain balance, frequency is inversely proportional to C or R. Frequency bridges of this type are made by General Radio and Muirhead. The chief disadvantage of a bridge for measuring frequency is the need for a detector sufficiently sensitive over the whole range of frequency covered and capable of rejecting any harmonics in the signal. For selective null indicators, see Sec. 5.23 (1).

Mention should be made of direct-indicating frequency meters. Several different types are available for accurately covering a small range, such as 45 to 55 c/s for giving warning of irregularities in mains frequency—very necessary in sound-recording etc. There are also wide-range valve-operated types in which the signal to be measured controls the generation of unidirectional pulses of the same frequency but constant duration. Their mean value, which can be read on a meter, is therefore proportional to frequency. The Airmec Type 265 works on this principle, covering 10 c/s to 100 kc/s in eight ranges. It is a portable all-transistor instrument, driven either by internal batteries or a.c. The pulse counter developed by the author primarily as a low-distortion f.m. receiver discriminator (*W.W.*, April 1956; also June 1956 and April 1958) could be used as a direct-reading frequency meter for 0 to 300 kc/s and could readily be adapted for other ranges.

The present trend, influenced by computor technique, is to measure both frequency and time by means of a digital counter. Frequency is indicated by counting the number of signal cycles in, say, 1 second. Time (such as the period of a low frequency) is indicated by counting the number of cycles of an internal oscillator occurring therein. High precision of the gating circuit that admits the signal for a specified time, and of the internal oscillator, is ensured by crystal control. The Marconi Instruments TF1345 valve-operated counter measures frequency from 10 c/s to 10 Mc/s, and the standard is a 100 kc/s crystal oscillator varying not more than $\pm$ 2 in 10^6 per week. Eight-decade display is by back-lighting numbered windows with miniature neon lamps. Using transistors and printed circuits, the Advance TC1 is limited to 1 Mc/s, but is only one-seventh the weight and one-thirteenth the bulk of the foregoing. When mains-operated, its accuracy is of the same order, but when battery-operated is considerably less because the crystal oven is out of action. The display is by six edgewise meters. A helpful feature of these counter-type instruments is that they can be used to test themselves. Although admirably versatile and accurate, their price range restricts their use to the more opulent establishments. The subject has been reviewed by P. R. Darrington (*W.W.*, June 1961).

Where frequencies to be measured are steady for at least a few seconds, one can make do with equipment available in almost every laboratory—calibrated signal generators and oscilloscope—as described in Sec. 10.10.

7.20. *Q* METERS

Owing to the difficulties of constructing bridges to give accurate results at r.f., resonance methods are often adopted instead. For some of these methods the necessary apparatus is assembled from instruments available, or extemporized to suit the particular requirements. But one important composite instrument must certainly be mentioned here—the Q meter. Although primarily designed for direct and rapid measurement of Q—this being perhaps the most generally useful figure of merit for tuning-circuit components—a Q meter can also be used for measuring r.f. inductance, capacitance, dielectric loss, transmission line parameters, etc.

For the meaning of Q, see Sec. 14.17, and for methods of use, Sec. 9.26. The principle of most Q meters, at least for frequencies lower than v.h.f., is shown in Fig. 7.19. A known r.f. voltage is provided by passing a standard current (monitored by the thermo-junction meter) through a low resistance r. This constant r.f. voltage is injected in series with the circuit being tested, which is adjusted to resonance, and the voltage across it is read by a valve voltmeter. A coil to be tested is tuned by a low-loss variable capacitor forming part of the instrument; a capacitor to be tested is connected in parallel with this standard and the resonant circuit completed by a low-loss inductor. On the assumption that Q is equal to the voltmeter voltage

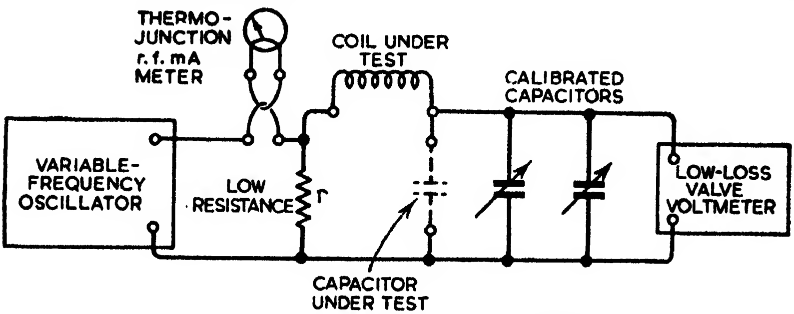


Fig. 7.19—Outline circuit of resistance-coupled Q meter. Instruments for higher frequencies use inductance in place of the low resistance r, *or alternatively a capacitance-coupling system*

divided by the injected voltage, the voltmeter is scaled to read Q directly. In practice, however, any coil has self-capacitance, and if this is an appreciable part of the whole tuning capacitance it is necessary to make a correction in order to arrive at the true Q. The accuracy of the direct readings is also affected by the series resistance r, the shunt resistance of the valve voltmeter, and series inductance of the instrument. If they are known, the readings can be corrected for them, but as this is a nuisance it is desirable that the instrument should be so constructed as to make the corrections negligible as often as possible. A typical value for r is 0·04 Ω, which can usually be neglected except in very high-Q circuits or above say 20 Mc/s. The loss due to the valve voltmeter should be less still. Instrument shunt capacitances can be incorporated in the calibration of the variable capacitor, which is preferably two capacitors—one main and another for reading differences of a few pF.

At very high frequencies, the smallest practicable r is likely to add appreciably to the resistance of the circuit being measured. In Q

meters for such frequencies it is therefore usually replaced by a reactance. Another questionable feature of the original instrument is the thermo-junction meter, which is fairly expensive and easily burnt out.

The Marconi Instruments TF1245 covers the remarkably wide frequency range of 1 kc/s to 200 Mc/s, but on the reasonable assump-

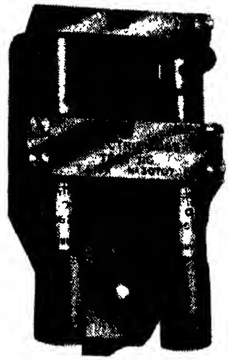


Fig. 7.20—Jig for adapting a Q meter (or other resonant measuring system) for tests on dielectric samples, which are placed between the disks on the right-hand micrometer. The left-hand micrometer controls a small increment capacitor. The system allows correction for edge effect. (*Marconi Instruments Ltd.*)

tion that not everyone will want to use it over the whole of this range the oscillator is not incorporated. Two oscillators, 40 kc/s to 50 Mc/s and 20 to 300 Mc/s respectively, have been designed for use with it; below 40 kc/s the lab's audio oscillator should do. For the v.h. frequencies a 20 mV input is developed across a 0·1 nH section of an inductive potential divider; for the lower frequencies, across a 0·02 Ω section of a resistive potential divider. The input across these dividers is measured at 0·5 V level with a crystal-rectifier voltmeter, in which special precautions have to be taken to avoid waveform errors—not a failing of the thermocouple type. A special low-inductance variable capacitor is used, and a disk-seal v.h.f. diode for the output voltmeter. Ancillary equipment includes the jig shown in Fig. 7.20, which facilitates use of the *Q* meter for permittivity and loss tests on dielectrics, and a set of inductors. For details of the TF1245 see *M.I.*, March 1959, and also the booklet *Measurements by Q Meter* (4th edn.), obtainable from the makers on request. It contains much general information on r.f. measurements and equipment, and also (like its previous editions) on the earlier model TF329G.

A relatively inexpensive *Q* meter by Advance (Model T2) nevertheless covers the useful frequency range 0·1 to 100 Mc/s. It follows the basic pattern of Fig. 7.19, except that the coupling is by reactance instead of resistance. A set of 12 individually calibrated coils, 0·1 μH to 30 mH, is obtainable for use with this instrument and as general laboratory standards of inductance.

A capacitive potential divider is used in a cheap and simple *Q* meter making use of a valve voltmeter (not necessarily low-loss) and an oscillator capable of an output of 1 V. It is due to J. Luijckx, who described it in *E.E.*, June 1957, p. 298. As Fig. 7.21 shows, it consists mainly of a double triode used as a pair of cathode followers, one for reading the input across the 100 : 1 divider and the other the output across the circuit being measured. If the former is set to 1 V, the latter multiplied by 100 is the Q. The cathode followers are first equalized by joining both their inputs to the oscillator. The success

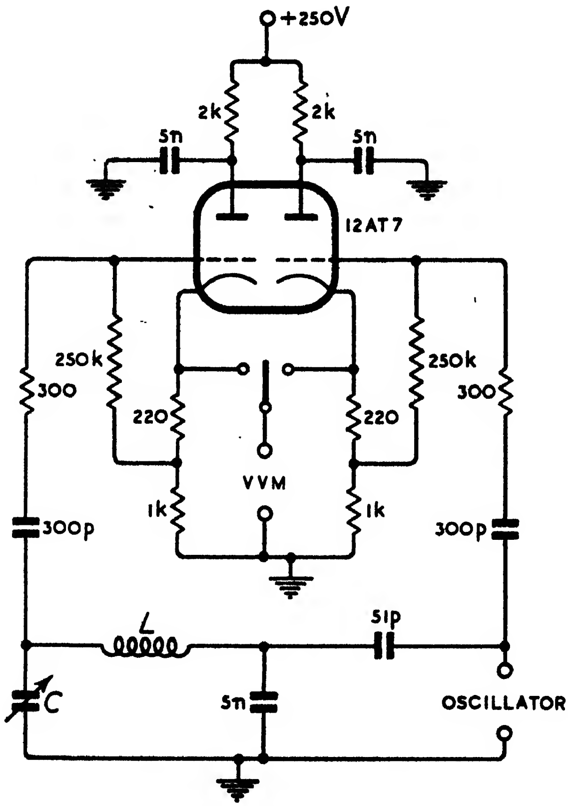


Fig. 7.21—Easily-made type of Q meter due to J. Luijckx, making use of an external r.f. oscillator. C_1 and C_2 form a 100 : 1 capacitive potential divider for feeding the resonant circuit

of this scheme would depend largely on the arrangement and quality of the components in the measurement circuit.

7.21. VALVE-TESTING EQUIPMENT

The number of valve types and the range of characteristics possessed by them are so enormous that any single valve tester is either very limited in its capabilities or extremely elaborate. Unless much money is available for comprehensive equipment it will probably be necessary to rig up for oneself something to suit the particular tests required, and information on such apparatus is included under Methods (Secs. 9.36 to 9.39). There is room here for no more than a brief survey of the different types of valve-testing equipment.

They may be divided broadly into two classes: those for laboratory measurement of characteristics, especially g_m, r_a and μ, under any desired conditions; and those for checking freedom from faults, such as short or open circuits, low emission and slope, softness of vacuum, etc. Although these latter are designed for radio servicing rather than measurement, there is surely no place where it is more desirable for valves to be free from faults than in the laboratory. Several makers produce valve testers in this class, and a question to ask when considering one is: Does it provide for connecting and feeding every type of valve liable to be used now and in the foreseeable future? It is usual to incorporate a vast selection of different valve sockets for coping with all the types our enterprising valve industry has produced, plus some device for connecting valves not yet imagined. One compromise is to have a single commonly-used type of socket on the tester, and adaptors for the rest. The provision of power supplies covering all possible combinations of requirements demands considerable ingenuity and is necessarily fairly elaborate, and moreover is liable to give rise to unwanted complications such as parasitic oscillation in the leads attached to high-slope valves.* Some of the problems (including that of expense) in providing power supplies are reduced in the Avo valve-characteristic meter by using a.c. only.

Valve-testing equipment, unless either used or designed with intelligence, provides exceptional opportunities for doing the wrong thing, to the detriment of valve, indicator or power source. In the Mullard tester these problems are dealt with by use of a cathode-ray indicator and a punched-card system to make the connections appropriate to the type under test.

Laboratory equipment proper includes at least three main types: (1) for plotting characteristic curves and deriving slope parameters from static or d.c. meter readings; (2) for a.c. measurement of slope characteristics, usually by a bridge or similar method; and (3) for displaying curves visually on a cathode-ray tube. To obtain anything like comprehensive valve data it is necessary to have both (1) and (2); (3) is more in the nature of a most desirable and useful luxury. The one necessity common to all is, of course, provision for feeding the valve electrodes at the desired voltages. A laboratory instrument has to have at least the adaptability of a servicing tester, plus means for measuring currents and voltages, especially in the static type, where high-quality meters are needed to obtain reliable results.

The VT 10 test set by A.P.T. Electronic Industries comprises (1) and (2), with stabilized power supplies and six meters (some of them multi-range). A useful feature is that the metered power supplies are available for external purposes. The same applies to another

* "Testing Steep-slope Valves", by J. C. Finlay. *W.W.*, March 1951.

instrument in the laboratory class—Metrix Type U-61B—in which parts (1) and (2) are obtainable separately.

Although cathode-ray equipment can hardly be expected to yield highly accurate quantitative results, it yields results in a minute fraction of the time, and enables one to see at once how the shapes of the curves change as adjustments are made. It also enables one to study the operation of the valve under dynamic conditions (e.g., oscillation), and to explore parts of valve characteristics where damage would be caused by static tests. An account of an elaborate equipment, is given by F. L. Hill and C. W. Brown of Cinema-Television (*E.E.*, Nov. 1949). A less complicated design is described by B. C. Foster (*E.E.*, May 1952). Still simpler apparatus that can be made up as required is shown in Sec. 9.37.

7.22. BRIDGES FOR VALVE CHARACTERISTICS

It is quite easy to work out balanced-signal methods for measuring the three basic parameters—amplification factor, μ; anode resistance, r_a; and mutual conductance, g_m—but even more than with circuit-impedance bridges a good deal of thought is needed to arrange the system so as to give the required results conveniently and with reasonable accuracy. There are of course the inevitable complications of feeding the valve electrodes and adjusting their voltages to the points at which the measurements are to be made—for although these parameters are sometimes called " constants " they are far from constant, and unless the precise working point is specified the figures have little meaning. In bridge circuits there is the additional complication of the stray admittances that the power sources may introduce. Means must be provided for balancing out valve interelectrode capacitances and other unavoidable reactance that would obscure balance. So equipment under this heading, too, is inclined to become rather elaborate.

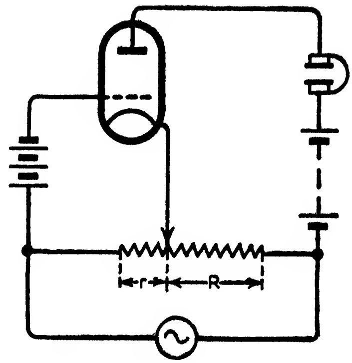
If a signal voltage in the anode circuit is balanced against a voltage in the grid circuit, the ratio of the two voltages is μ. Fig. 7.22 shows in principle a very simple method of supplying and adjusting the two opposing voltages. When the tapping (or the resistance each side of it) has been adjusted to give silence in the phones, μ is given by R/r. Fig. 7.23 is an elaboration of this circuit to include an adjustment (C) for balancing out reactance. A convenient R is 1 kΩ; with r, say 0 to 500 Ω, calibrated to read μ directly. To avoid excessive crowding of the high values the rheostat should be " tapered " to give relatively slow variation near the zero end.

By switching in the resistance R_1 the same circuit can be used for measuring r_a. When balance is again achieved,

$$r_a = R_1 \left(\frac{r}{R} \mu - 1 \right).$$

It will be noticed that if r and R are the correct values to balance for μ, it is impossible to balance r_a. But if, when r and R are right for μ balance, a resistance $R_2 = R/101$ is switched in at the same time as R_1,

Fig. 7.22 (right)—Outline circuit showing principle of μ bridge. In practice, especially with high-μ valves, it is necessary to balance out stray capacitance, as shown in Fig. 7.13



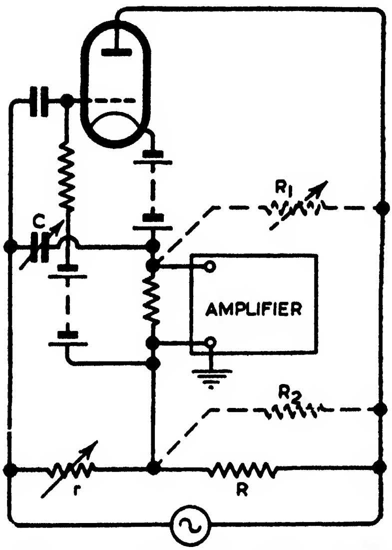


Fig. 7.23—Elaborated valve bridge, showing capacitance balance, amplifier for increasing sensitiveness of detecting balance, and additional arms for measuring r_a

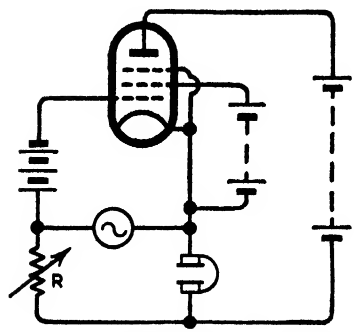


Fig. 7.24—Simple form of circuit for measuring g_m

the formula simplifies to $r_a = 100R_1$, and this is very convenient for making R_1 a direct-reading control.

When μ and r_a have been obtained in this way, g_m follows from $g_m = \mu/r_a$. But it is advisable to have means of measuring g_m directly, as a cross-check, and because it is the most important of the three as a guide to the condition of the valve. Fig. 7.24 shows a circuit for doing this, and the calculation is simple: $g_m = 1/R$. It would be possible, without unduly complicated switching, to combine this apparatus with Fig. 7.23.

In all these arrangements, both signal voltages are obtained by potential-dividing a single source, but this simplification creates difficulties in the arrangement of power supplies. E.g., in Fig. 7.24 three separate batteries are needed, and it would be impracticable to use mains power units because of stray admittance. Details of a pair of bridges (one for μ and r_a; the other, which can be combined with it, for g_m) suitable for mains drive (except for grid bias) are given by G. Smith (*E.E.*, March 1952).

The most flexible method, giving freedom to use conventional power feeds, and avoiding errors due to signal-source resistance, is to inject the required signal voltages wherever required, from separate transformer windings. It was developed by W. N. Tuttle (*Proc. I.R.E.*, June 1933), as a basis for the General Radio valve bridge. A recent design for very high accuracy—of the order of 0·1%—was described by M. R. Child and D. J. Sargent (*Proc. I.E.E.*, May 1959). Such accuracy is much beyond what is needed for ordinary tests, and calls for very precise adjustment of supply voltages; in this case the accuracy was needed for detecting deterioration on life test. The circuit used for g_m is basically Fig. 7.24, and for anode conductance ($1/r_a$) the input signal is transferred from grid to anode. Measurement of g_2 amplification factor is also provided.

7.23. TRANSISTOR TESTERS

Although transistors do not run to so many electrode combinations as valves, they are inherently more complicated because the action between input and output is both-way. So four parameters are needed instead of two. Moreover these are more frequency-dependent than in valves. And the testing of both *pnp* and *npn* types has to be provided for. But the innumerable holder problem is absent!

As with valves, transistor testers can be divided broadly into two classes: those for checking that a transistor is not faulty and for getting a general idea of its more important properties, and those for precise and comprehensive measurements. Since manufacturing tolerances are so wide, the latter are seldom needed outside the laboratories of transistor producers. It is not difficult to make up a simple instrument in the former class, but some of those available commercially are so cheap that it is hardly worth while doing so.

An example is the Ediswan R.2285 " Beta Tester ". The transistor under test is connected as an oscillator, and a gain control is adjusted to the threshold point of oscillation, indicated audibly. The gain control is calibrated directly in current amplification factor in the common-emitter configuration, usually denoted by β or α', though α_e is a more logical symbol. A meter reads collector current for the β test on a 0 to 5 mA range, and on a 0 to 0·5 mA range the " leakage current " I'_{co} (or I_{coe}) which flows when the base is open-circuited. This is a good clue to the quality of the transistor, and normally should be not more than a few microamps. Three quick-release terminals are provided for connecting the transistor.

The Advance Type TT.1 provides similar facilities, with the considerable advantage that the transistors can be tested *in situ*.

The Cossor Model 1325 is a little more elaborate. The power supplies are derived from a.c. mains, and, besides I_{coe} and α_e, the collector turnover voltage—the voltage at which the current with open-circuited base begins to rise steeply—can be measured, and *npn* as well as *pnp* transistors are provided for. The measurements are purely d.c., by the method shown in Fig. 7.25. With the switch open, I_{coe} is read.

Closing the switch enables I_{coe} to be balanced out; a known base current is then supplied and the reading again noted. It represents the increase due to the base current, and the ratio of the one to the

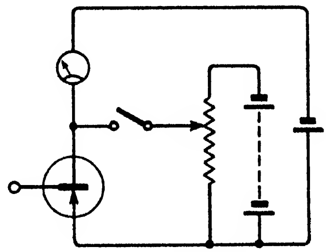


Fig. 7.25—Basis of d.c. method of measuring α_e and I_{coe} employed in the Cossor 1325 transistor tester

other is the large-signal value of α_e. The turnover voltage is measured directly between collector and emitter terminals with base open-circuited and a known high voltage (65 or 85) applied through a known resistance (50 kΩ or 10 kΩ). The meter is protected from overload by the rectifier shunt described in Sec. 5.5.

Hatfield Instruments recognize that many experimenters do not care to tie up a lot of money in a transistor tester, so they have worked out a scheme to enable the hybrid or *h* parameters to be measured by means of their 15 c/s to 1 Mc/s millivoltmeter, an a.f. oscillator, and suitable power supplies. (The four *h* parameters specify the transistor as a " black box " or quadripole. They are, of course, complex quantities when measured with a.c.; that is, they have both magnitude and phase.) Fig. 7.26 shows their recommended circuit, which as one can see, is simple and inexpensive. Even more so is a ready-made instrument by Grundy & Partners, arranged for clipping directly on to an Avometer or similar multi-range meter. Both *pnp* and *npn* transistors can be tested for α_e, I_{coe} and I_{cob}, and there is also provision for testing diodes up to 10mA forward current and 9V reverse potential.

Where transistors are used for r.f. or pulse work it is most desirable to be able to measure their performance at high frequencies; at least f_α which is the frequency at which α is reduced by 3 dB, and preferably the four admittance parameters. This can be done with the Wayne Kerr r.f. bridge B.601 already mentioned, in conjunction with a control box and a set of adapters, of which one set is available for *pnp* transistors and another for *npn*.

For fairly detailed examination of transistors in a wide range of types, the Avo transistor analyser is convenient. It is battery operated, and covers *pnp* and *npn* types of low and high power. D.c. characteristics, turnover voltage, I_{coe} down to 2 μA, α_e at 1 kc/s and 0·5 μA input, and noise are all measurable. Provision is made for testing

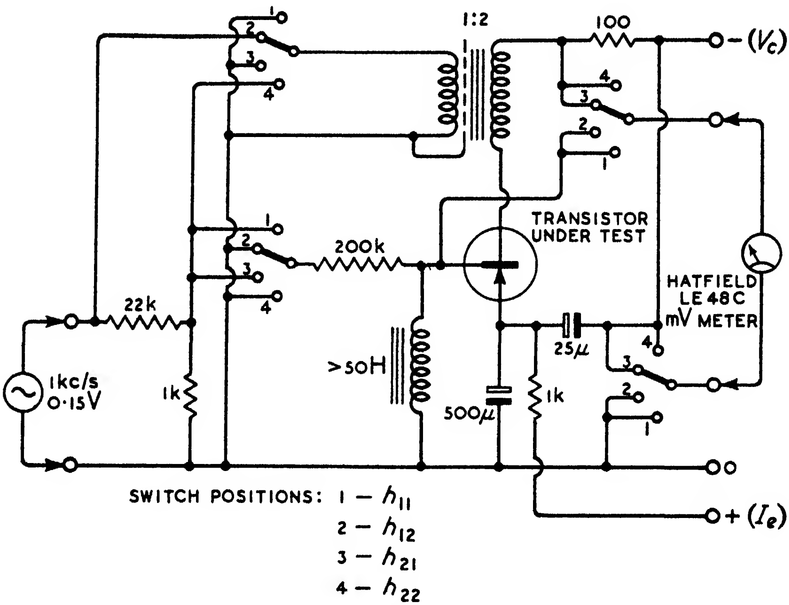


Fig. 7.26—Hatfield Instruments circuit for measuring transistor hybrid (" h ") *parameters*

transistors *in situ*; the meter, oscillator and ×5,000 transistor amplifier being available for external use.

The following articles describe methods and apparatus for the various kinds of tests and measurements just outlined. A simple test set for measuring I_{coe} and α_e with d.c., somewhat as in the Cossor instrument, is due to J. N. Prewett (*W.W.*, Aug. 1958). An OC76 transistor emitter-follower is used as a low-resistance variable-voltage tapping in the manner explained for the analogous cathode-follower in Sec. 4.8. A test set by G. G. Yates (*E.E.*, Oct. 1959) also measures I_{coe} and α_e with d.c., but is provided with a transistor amplifier so that with the aid of the signal generator and valve voltmeter that even a modest laboratory possesses it measures f_α and a.c. input resistance with short-circuited output (h'_{11}, one of the " hybrid " parameters). R. Hutchins and J. D. Martin (*E. & R.E.*, Oct. 1959) explain how to measure all four hybrid parameters, using simple bridge methods. An alternative to these or other sets of input, output and transfer parameters is the specification of a transistor in terms of an equivalent circuit, of which the T form is probably the most used.

A set for measuring the elements of this circuit at 1 kc/s is fully described by R. A. Hall (*E.E.*, Feb. 1958). The ability of the transformer ratio-arm bridge to measure the direct impedance between two points while ignoring the admittances between them and a third point (such as earth) is applied to transistors in a good practical article by M. J. Gay (*B.C. & E.*, June 1959). The same author contributed a paper on the measurement of small-signal transistor parameters to *Proc. I.E.E.*, Vol. 106, Part B., Supp. 15, May 1959, where also H. G. Bassett described apparatus for α_e and α_b over the wide frequency range 1 to 300 Mc/s by signal injection and measurement via wide-band transformers. C. Bayley gives particulars of equipment that can be made up for measuring most of the useful transistor parameters (*W.W.*, Aug. 1961).

Lastly, J. F. Young (*E.E.*, June 1959) describes in some detail a set for displaying transistor or diode curves on a separate oscilloscope. A Dekatron is used to develop a stepped voltage controlling the base current. At each step a half-sinusoidal voltage is applied to the transistor and the resulting collector current is plotted against voltage.

7.24. RECEIVER ALIGNMENT TEST SETS

For aligning the tuning circuits of receivers an almost indispensable aid is a "wobbulator" or frequency-swept signal generator in conjunction with an oscilloscope—see Secs. 4.36 and 5.38. Although the equipment can be assembled from separate units, sufficiently constant use would justify one of the composite sets designed for the purpose. In the " lab " class there is the Marconi Instruments TF1104, which comprises a signal generator covering the v.f., i.f. and v.h.f. bands used for television and f.m. broadcasting, an oscilloscope with time base synchronized with the signal frequency sweep, and a crystal-controlled oscillator for providing accurate frequency calibration marks on the trace. The sweep width is adjustable from 0·5 to 10 Mc/s.

Designed mainly for servicing receivers, and therefore less expensive and less precise but even more versatile, the Airmec " Radivet " and " Televet " have many obvious uses in the laboratory too. The Radivet comprises an a.m./f.m. signal generator with crystal calibration, an a.f. oscillator and an oscilloscope, which can be used separately or in conjunction. Besides alignment of tuning circuits and f.m. discriminators, a.f. distortion can be studied, and frequencies and voltages measured, to mention only a few uses. The Televet performs similar services especially on the television frequencies, and in addition includes a pattern generator and e.h.t. voltmeter (using the c.r.t. as indicator).

7.25. DISTORTION METERS

A considerable choice of equipment is available for measuring distortion, but because its design is very much bound up with the nature of distortion and the principles of its measurement, the question of equipment is discussed along with these other aspects, in Secs. 11.15–16. Commercial wave analysers and distortion-factor meters have been reviewed by R. Brown (*B.C. & E.*, Oct. 1960).

CHAPTER 8

Choice and Care of Equipment

8.1. GENERAL POLICY IN CHOOSING EQUIPMENT

The four previous chapters, if read straight off, may have had a slightly bewildering effect on anyone faced with the task of equipping a laboratory with limited means, in spite of the fact that elaborate and expensive apparatus has either been omitted or, if of general importance, described quite briefly. There is a vast amount of equipment that might have been mentioned, but has been left out as too highly specialized or likely to be available only in lavishly appointed laboratories. Even so, much more has been included than most people are likely to be able to acquire personally.

Of course, it is no good publishing a list and saying: " This is what you should have ", without knowing what it is needed for and how much can be spent on it. Nevertheless it is possible to offer some general advice and suggestions for particular items that should be included in any laboratory intended to carry out measurements and experiments in radio, electronics, etc. Some of the general advice was included in Chapter 1; now, in the light of the intervening chapters, we can take up the subject again.

One method is to make a list of all the kinds and ranges of measurements that are likely to be required, and then decide on the equipment needed to cover these most economically. Like J. K. Jerome's three men planning to spend their holiday in a rowing boat, who made a list of the equipment that was absolutely essential and then found that the Thames was not navigable by a vessel sufficiently large to contain it, one may have to tear up the first list and start again with a fresh interpretation of the word " essential ". Few, if any, investigators can afford to obtain the most suitable equipment for every job they undertake. The advantages of having exactly the right equipment are that it enables the work to be done accurately in the shortest possible time, and it does not demand ingenuity and thought to extemporize a method using something less suitable. But if one did have everything needed, it would all have to be housed and looked after, and the more there was of it the less often each instrument would be used and the less familiar one would be with its operation and condition. Such an imaginary " ideal " laboratory would not then be an unmixed blessing. There is a good deal to be said for having a few reliable and well-understood instruments and a sound basic theoretical knowledge (tempered by practical sense) to enable one to devise methods making use of what there is, supplemented by apparatus that can be rigged up for particular needs. The following quotation from F. E. Terman can hardly be over-emphasized: " Success in making measurements in radio work is primarily a matter of having available a satisfactory technique that is thoroughly understood rather than of having available innumerable alternatives."

It is wise, then, when equipping one's laboratory to give preference to instruments of lasting value and wide application.

8.2. A SHORT LIST

To get down to something definite, the following are recommended as worthy of first consideration in a laboratory aiming at tackling as much as possible in the most economical way.

(1) *Stabilized power unit* (Secs. 4.3 to 4.7) to give up to not less than 100 mA at voltage variable over the range of interest (e.g., 150 to 450), with at least one cathode-follower tapping (Sec. 4.8) and the usual a.c. supplies for valve heaters. Where stabilization is not essential, but variable output voltage is, the cathode-follower rectifier type of unit referred to on p. 35 is recommended. For transistors, if the fixed voltages of batteries are not sufficient, a variable-voltage stabilized unit will be needed, with current output up to at least 1 A.

(2) *A.f. signal generator*, such as the type described in Sec. 4.28. It should certainly be calibrated in frequency, as it can then be used for frequency measurement as well as a signal source. A calibrated attenuator is a most valuable facility.

(3) *R.f. signal generator* (Sec. 4.30), with a specification to cover as many of one's needs as can be afforded. For the same reason as (2) a reasonably accurate frequency calibration is the first requirement, even if it is only a home-made oscillator or inexpensive " servicing " signal generator. Output control and calibration are the next priorities.

(4) A good *multi-range meter*, including ohmmeter; and by " good " is meant not only the number of ranges the maker can advertise but its reliability (Secs. 5.1 to 5.8). This instrument is so much used that it is worth getting one that gives ample facilities and will stand up to years of hard work.

(5) A *cathode-ray oscilloscope* (Secs. 5.25 to 5.41). Something under this heading is a necessity, even if one can only afford a second-hand tube with a home-made power unit. Choice of an oscilloscope needs careful consideration, for the more expensive of two is not necessarily the better for the purpose. With suitable amplitude control and provision for comparison with a known signal, a c.r.o. may render a valve voltmeter unnecessary.

(6) *Valve voltmeter.* This is put down here for consideration, not necessarily for acquisition. Although certainly an important and useful class of instrument, it is not so indispensable as sometimes believed. Much of the work it used to do can be done by a cathode-ray tube, especially a flat-screen type with associated calibrating facilities. And choice from among the different types of valve voltmeter is so complicated that it can hardly be made apart from a knowledge of the work in view. It is possible to combine (4) and (6) by getting an "electronic test meter", but this usually necessitates a mains connection (which may not always be available) and a warming-up delay, so although it provides more facilities it is not always to convenient to use.

(7) *Decade resistance box* (Sec. 6.3). At least one, up to 11,110 Ω in steps of 1 Ω, is practically indispensable where anything like accurate measurements are to be made. In fact, both for measurement and general experimental work there can never be too many! But for the less exact purposes one generally has to make do with calibrated rheostats and switched selections of "preferred value" resistors, which should be provided on a generous scale.

(8) *Calibrated variable air capacitor* (Sec. 6.7). This is needed as a standard of reactance, in bridges and for r.f. measurements among other things. A 1-nF (1,000 pF) model, of as high quality as can be afforded, supplemented by one for low capacitance, say 1 to 20 pF, which can be calibrated from it (Sec. 9.34), covers most requirements.

(9) A *C-and-R bridge* in the class described in Sec. 7.14 is the most convenient and economical instrument for checking component values within a few per cent.

(10) *Apparatus for more accurate measurement of R, C, L, Q, etc.* This is another of the departments where recommendation without knowledge of needs becomes difficult. Bridges to cover all possible ranges of everything with high accuracy are outside moderate means, but a "universal" bridge in the 1% class may be considered; failing which it is a case of seeing whether one can make do with (9) (possibly in a higher-quality version) plus something like the Turner-Owen bridge (Sec. 7.16) for inductance. R.f. coils are the chief problem, because they usually have to be measured rather accurately. Measurement at r.f. by resonance methods can be done quite accurately *and* cheaply if one is prepared to take time and trouble over it (Secs. 7.20 and 9.25 to 9.35). A set of plug-in inductors, preferably measured for L and Q, is a necessary adjunct for some measurements.

In addition to anything of a more specialized nature that should come into the first-priority list because of one's own special line of work, there will be of course a few single-range meters, rough resistance and capacitance boxes, heater transformers, and miscellaneous components.

No valve tester has been included in the list. The difficulty is that either the latest sort of valve that one wants to test seems somehow to be outside the scope of the instrument, or else, to avoid this possibility, the designer has made the apparatus so flexible that it is almost as much trouble to make all the appropriate connections and switchings as to rig up a test for oneself from the appropriate valve socket and meters. This being so, the author's practice when setting up an experimental circuit involving valves is to check the valves in their own circuits, and, on the comparatively rare occasions when a valve is being investigated as such, to hook up a suitable circuit. Much the same holds good for transistors. Admittedly this is not likely to suit workers handling very many valves or transistors, for whom proper equipment (Secs. 7.21–23) may be justifiable.

8.3. DEVISING SPECIAL APPARATUS

That completes our present review of the sort of equipment that is most practical for general purposes; there is endless scope for individual ingenuity in adapting such means as are available to meet the task in hand. The more one follows the line of original research the more it becomes necessary to devise special instruments for carrying it on. The keen experimenter should take every opportunity of studying the scientific papers and experiments of experienced workers, noting the ways in which investigations of great beauty and precision are sometimes made with very simple means. Many of the fundamental laws of electricity were established in this way. The work has become so highly developed that nowadays very great instrumental resources are necessary to carry on much of it, but simple methods directed by genius are still powerful.

If the available instruments cannot be adapted to the job, it may be that the job can be adapted to the instruments. Measurements on the radiation patterns of aerials, that might be impracticable or at least inconvenient, can be managed by reducing all dimensions (including of course the wavelengths of the signals) on a suitable scale. And if the frequency range of an oscillator fails to cover the whole band over which one wants to test a filter design, the test can perhaps be performed on a filter in which, say, all the capacitances have been halved or doubled and the input and output resistances altered to suit.

8.4. PREPARATION FOR AN EXPERIMENT

Except for routine work, and particularly before starting an investigation that may absorb much time, it is a good idea to run through *E.T.* (late *W.E.*) annual indexes of abstracts and references or *Science Abstracts* for some years back and make a note of any publications that might have a bearing on the matter, and then look these up. They might save a lot of time. But while it is a good thing not to duplicate other people's work unnecessarily, it is also a good thing not to take the correctness of their results too much for granted, nor to be deterred by any lack of success they may have experienced.

Then in setting up what may perhaps be an elaborate assembly of apparatus, the result of hurrying over the preparations to get down to the work itself is likely to be the same as that of rushing straight into an examination paper without calmly pausing to study the questions—misdirected effort and irrelevant answers. The business of thinking out and arranging the apparatus should be separate from the business of using it, so that one can concentrate on each in turn. Do the readings it is proposed to take really supply the required information? Has the method been designed so that the results are given in terms of data that are the most reliable available? (Sec. 3.11.) Are any instrument calibrations that are to be used dependable, and the instruments themselves in good order, with clean contacts and firm connections? Is it possible to check the results, by arriving at them by at least one other independent route? Is the apparatus arranged so that readings can be taken with the minimum of alterations and adjustments, such as might introduce errors? When many different readings have to be taken, it may be worth while running a "pilot" test—covering the whole ground quite quickly, with readings perhaps more widely spaced and taken without regard for accuracy, and plotted roughly as a graph. If instead one starts taking very carefully the full complement of readings, it may be found that an instrument reaches the limit of its range before they are complete, or some other snag may arise to necessitate a modification of the procedure and a fresh start.

In spite of the most careful planning, things sometimes go wrong; for example, a result obtained on one occasion fails to repeat itself on another occasion when the conditions are apparently similar. It may happen in the course of a single experiment, especially with complicated apparatus. It often happens when a system has been worked out experimentally, and is then put into proper shape for permanent use. This sort of difficulty is much more easily cleared up if the original apparatus has been left intact, although even then it may be very puzzling. Without it, the secret may be lost for ever. It is necessary gradually to eliminate the differences between the two, until the one responsible for the unexpected divergence in results is revealed. Often, as in detective fiction, the offender turns out to be the least suspicious character. A method of minimizing this problem is to transfer from breadboard to final form in stages, checking the performance of the whole as each stage is completed. For the question of dealing with results after they have been obtained, see Chapter 13.

8.5. IMPORTANCE OF HANDINESS OF INSTRUMENTS

If much work or thought is required to bring certain apparatus into use there is a tendency not to bother to make the measurement at all. Inconvenient instruments are simply not used. On the contrary, if instruments are chosen and arranged to be really *handy*, it is easy to develop the habit of checking everything and thus avoiding much subsequent loss of time and perplexity. The author found that he was getting into a habit of using one particular meter in preference to others, not entirely because it included a particularly useful selection of ranges *but just because it happened to be fitted with clip leads*, thus requiring less effort to pick up and apply than those other instruments for which one had to look about for a pair of leads and connect to terminals. Incidentally, it is easier to see what one is doing and harder to make mistakes with a plug-in system of range-changing than with switches. Too much emphasis cannot be laid on the value of a good stock of crocodile-clip leads, preferably painted with some sort of enamel to minimize risk of short circuits. In some laboratories one knows, whenever anything is taken in to be examined there is a search for wretched bits of wire to twist around soldering tags, giving thoroughly unreliable connections everywhere.

8.6. LAYOUT OF APPARATUS

Then the layout of the apparatus is often important. There are some connecting leads that could be taken round the town and back without impairing results, and others where every inch is vital. In v.h.f. work, the standard of impedance may be an inch or so of wire. By giving thought to the placing of the instruments, the leads likely to cause stray couplings or undesired distributed impedances are reduced to a minimum, and so are couplings between such coils as it may not be possible to screen effectively. Some parts of the circuit ought, perhaps, to be kept as far away from material substances as possible, or from other parts in which currents of the same frequency but a very different power level are flowing.

Particular care is necessary to avoid trouble due to stray fields. Readings taken with moving-iron meters are easily vitiated by magnetic fields. Hum from the mains readily intrudes by way of either magnetic or electric field, and occasionally by conductive leakage. Thought must be given to earthing and screening, in order to minimize electric fields and to bring stray capacitances across parts of the circuit where they do not matter (Sec. 7.6). Magnetic fields of radio frequency can generally be excluded by copper or aluminium screens (not necessarily earthed) with good conductive paths around the magnetic flux; but for power frequencies mumetal is the suitable material.

8.7. NEED FOR OBSERVING RESTRICTIONS

It is very important to be fully aware of what is inside each built-up unit. Failure to do this may result in short-circuits (because terminals thought to be unconnected are " commoned " or earthed), or omission of connections (because terminals thought to be linked are actually unconnected), or overloading some instrument, or exceeding the conditions within which calibration holds good. For example, an output meter, sold as accurate within a certain limit of error, may be subject to much larger errors if an excessive standing current is passed through the transformer. Or a circuit may be effectively by-passed only for a certain range of frequency, and error caused by using it outside this range; or the stray reactances may be greater than is allowable for the work intended. These risks are greatest with apparatus of neat " commercial " appearance, in which the convenience of operation and the engraved wording may deceive one into thinking it *must* give the correct results regardless of the many assumptions that are tacit unless a detailed book of conditions and exceptions is provided and consulted. A breadboard is at least easy to follow, and a sizzling coil is spotted sooner than if it is in a substantial teak cabinet. All apparatus, particularly of the more specialized types, should have a circuit diagram and a summary of conditions of use pasted in or on it, or at least readily available in the loose-leaf instrument book recommended in Sec. 8.11. An idea that should be more widely used is that adopted by Cossor in their instruction books—one circuit diagram permanently attached, and a duplicate for tearing out to keep with the instrument.

8.8. PERSONAL RISKS

Some technicians seem to consider it clever to affect a contempt for electric shock. One wonders how many of the fatalities are pure accidents and how many arise from carelessness. Sometimes a shock of thousands of volts causes only minor discomfort, and sometimes a hundred volts or even less is fatal. While most modern television-receiver e.h.t. generators are not lethal, it is wise not to take chances. In any case, almost any ordinary a.c. domestic receiver gives 1,000 V peak across the transformer secondary supplying the valve h.t. It is difficult to get a dangerous shock when standing on an insulating mat with one hand in the pocket, even when poking about among high-tension apparatus; not that it is wise to do so in any circumstances, but acquiring a one-hand habit might save your life. Of course, one ought never to touch conductors for the purpose of ascertaining whether or not they are " live ", but in an emergency when no instrument is available the rule goes by the board. When it does, at least make sure that you use the *back* of your finger, so that any resulting muscular twitch breaks the circuit instead of increasing the contact pressure. If contact is made *very* lightly it is not only relatively safe but is extremely sensitive (to A.V.).

8.9. AVOIDING DAMAGE TO INSTRUMENTS

Serious personal harm is fortunately rare, but the involuntary jerk resulting from even a slight " packet " often damages some delicate instrument or causes short-circuits. This is only one of very many possible causes of damage to apparatus. It is so very easy to ruin a valuable instrument by a moment's inattention or lack of foresight; and when a whole assembly of them is being used in one experiment the risk is so much greater. It is possible to do damage without being aware of it; for example, in the preoccupation of carrying out a series of readings one may not realize that the current passing through a decade resistance box has risen so high as to produce a permanent loss of accuracy. It is therefore particularly important that the standards which are ultimately relied upon and which there may be no means of checking are exposed as little as possible to such risks. Though an instrument may appear to have recovered from an overload, one always has uncomfortable doubts about the accuracy. Thermocouple meters are the most vulnerable to comparatively small overloads, and replacement and recalibration is expensive. Protective devices for meters have been given a good deal of attention lately, but the presence of fuses or cut-outs should not be made an excuse for carelessness. In this book strong preference is given to instruments and methods that relieve one of all anxiety concerning safety. An example of a method that is *not* recommended is the measurement of capacitor leakage by connecting the capacitor in series with a delicate microammeter and a high voltage. Even if one takes the elementary precaution of short-circuiting the microammeter until the charging current has passed, the certainty of disaster in the event of capacitor breakdown puts the method out of consideration. Sometimes accidental short-circuits do the damage; sometimes open-circuits, as when the bias comes off the grid of a valve, in the anode circuit of which is a delicate instrument with the standing current carefully balanced out. These balances are especially risky. The method of resistance measurement recommended in Sec. 9.1 has the great merit that although a sensitive instrument is used, no resistance from zero to infinity inclusive can drive it off the scale on any of its ranges.

When such fool-proof devices cannot be used, make a calculation beforehand of the maximum current that can flow anywhere it might do harm (Fig. 14.2 helps with this); make all connections secure; avoid exposed leads or connections that might fall or be drawn into contact; and before operating switches make quite certain that they are going to do what you expect. It is an excellent habit to leave all multi-range meters connected for the highest volt range when they are not otherwise being used. This should be done immediately after a reading is taken, even if it is wanted again in half a minute. It might happen to be wanted to read volts instead of milliamps!

Another good habit is to short-circuit the terminals of meters, especially the more delicate types, before moving them. The heavy damping caused by the generated current cushions the moving parts. And avoid shock to the instrument, for its jewelled bearings are more easily damaged than is generally realized. The more precise the instrument, the greater the risk, for to reduce friction to a minimum the size of the bearing surface is made so small as to impose nearly the safe limit of stress on even steel and jewel.

It is not only during actual use that harm can befall instruments. Many of them are liable to be affected by damp, fumes, extremes of temperature, dust, and bright sunlight—roughly in that order of importance. Modern design and materials are tending to reduce trouble from these causes, but in any case it is desirable to keep as even a temperature and humidity as possible where accurate work is to be done, to preserve calibrations. Unless the temperature is kept above the surroundings in cold damp weather there is almost bound to be trouble from condensation of moisture, drastically reducing insulation resistance and corroding fine wires. The lab. heating should not itself produce water vapour, and ventilation is desirable to carry off moisture (Sec. 2.3).

8.10. MAINTENANCE: CONTACTS

It is easy to advise other people to carry out regular maintenance of equipment, but quite another matter to do it oneself. Provided that it is kept clean and dry and is not misused, most electronic lab. equipment needs very little maintenance. Exceptions to this rule should be avoided!

But switches and other contacts need occasional attention if their resistance is to be kept low and constant. There are two main types of contact: those between metals such as brass, copper, and bronze, which quickly tarnish, and need fairly heavy contact pressure to break through it; and precious metals, which give low-resistance contacts with comparatively light pressure and (being very thin) would be injured by heavy pressure. Silver occupies an intermediate position.

If the contacts are hermetically sealed in dust-free compartments, precious metal contacts should keep good indefinitely, and base metals need only an occasional operation to clean them when they have been out of use. But if dust is allowed to accumulate they become rough and their resistance fluctuates. Provided that this stage is not allowed to develop far, it should be enough to wash away dirt-retaining grease with a suitable solvent such as trichlorethylene, used sparingly. Carbon tetrachloride is often used, but the commercial fluid is likely to contain traces of free acid, bringing risk of corrosion. If liquid cleaning by itself is not sufficient, a slip of rouge paper folded to present an active surface on both sides should first be carefully drawn between the closed switch contacts. On no account should even fine emery be used, unless the contacts are seriously rough or pitted, and then care must be taken to keep flat surfaces from being rounded off or misaligned (back the paper with a steel rule and work in parallel strokes) and to remove all traces of abrasive by cleaning. If contacts plated with silver or gold are cleaned harshly they will be ruined. After cleaning, the contacts should not be left dry but be given a *trace* of lubricant.* This can conveniently be applied by dissolving

* "Cleaning Switch Contacts", by J. J. Payne. *W.W.*, Feb. 1948.

a small proportion of Vaseline in the cleaning fluid, to provide a thin film on the contacts after the solvent has evaporated. When used, care must be taken not to slop it all over the apparatus, where it might reduce insulation by attracting dust, etc. And of course never use fluffy material for cleaning. The sliding contacts of rheostats, etc., can be similarly treated.

A proprietary switch-cleaning fluid is prepared by Servisol Ltd., 14, North John Street, Liverpool, 2.

8.11. CALIBRATIONS

The other thing that should be attended to is occasional checking of calibrations. The frequency with which this should be done increases steeply with the number of people who use the instrument. *Results of all calibrations, modifications, etc., should be recorded.* No amount of personal inertia or pressure of business must be allowed to delay the instituting and maintaining of a handy loose-leaf book containing a record of every significant instrument. Each should have a serial number painted on it in a distinctive colour, corresponding to its number in the book. Incidentally, these serial numbers greatly reduce the task of writing up experiments, being briefer than a description and more easily referred to than the makers' numbers. The book record should include in concise form a specification, circuit diagram, calibration curves, operating details, precautions, and a history of the instrument from its source, including recalibrations, modifications, and in fact any data likely to be needed during use or that might concern results obtained. Voluminous data not of general use should be kept elsewhere, with a reference in the book. The time spent on keeping the book up to date is well worth while, and might be exceeded by time spent in tracing the necessary information without it, or obtaining wrong results through ignorance. It is very satisfactory, when one comes to use an instrument (having forgotten nearly all about it) to note its number, turn it up in the book, and find everything one wants to know. Although it is a good principle to have at least the circuit diagram attached to the instrument (just in case the precious book should ever be mislaid) this is not always practicable. The results of recalibrating instruments must usually be recorded as error curves. The inconvenience and possibility of slips in using these can be minimized by putting them into a suitable form. This art has been discussed elsewhere by the author (*W.W.*, May 1954).

Before an attempt is made to calibrate or recalibrate an instrument, it is worth while checking that it is in a fit state. If a meter has excessively loose or tight bearings, or invisible hairs on the scale plate to impede the pointer, or a capacitor is mechanically unable to retain a calibration, or anything has loose connections or high-resistance switch contacts, then something should be done about it before spending time on calibration. Meters should be tested by raising and lowering their current slowly and steadily between zero and full-scale, watching carefully for any signs of stickiness. Another test is to use a rheostat that is variable in small steps to note whether the pointer returns to exactly the same reading from positions slightly below and above. Few instruments are perfect in this respect; the others need *gentle* tapping before taking a reading. If ever it is necessary to uncover a meter, it should be done in as clean an atmosphere as possible, and great care taken not to allow any dust or fluff (still less iron filings!) to gain admittance. If when in normal use a meter has a metal case or is mounted on a metal panel, its readings may be appreciably different when these are not in position—a possibility that must be kept in mind when adjusting shunts or multipliers. Beware of stray fields! The author once wasted the best part of a day tracing the cause of erratic readings (which turned out to be the magnetic field due to a tubular rheostat affecting moving-iron meters) and repeating a long set of calibrations that had been invalidated thereby. And remember that when even a good moving-iron meter is used on d.c. the readings may differ slightly according to the direction of current. Although moving-coil meters usually have linear scales, in accurate work it is not safe to assume perfect linearity.

Finally, in any calibration make sure that the data on which it is based are reliable and that there are no abnormal conditions to vitiate results. Once errors have crept in where they are least likely to be expected, there is no knowing where they will end.

Ref: " A R.A.F. Calibration Centre ", by W. H. Ward and others. *Proc. I.E.E.*, Part III, Jan. 1950.

CHAPTER 9

Measurement of Circuit Parameters

(A) MEASUREMENT AT ZERO AND AUDIO FREQUENCIES

9.1. RESISTANCE: MEDIUM VALUES

For accuracy within a few per cent at z.f., the direct-reading ohmmeters described in Sec. 5.8 and included among the facilities of multi-range test sets are the most convenient. They have the advantage over bridges that they can be used even when the resistance being measured is varying. The general method of use, given in Sec. 5.8, consists of a preliminary zero adjustment with the terminals shorted, and then a reading with the resistance to be measured (call it R_x) connected.

If no ohmmeter is available, a voltmeter can be used. The nearer the resistance of the voltmeter (R_m) is to R_x the better the accuracy. The meter is first used to measure the voltage V_1 of a suitable source, whose resistance must be negligible compared with that of the voltmeter. V_1 is preferably at or near full-scale. Next, the measurement is repeated with R_x in series, as in Fig. 9.1, the reading now being V_2. Then

$$R_x = R_m\left(\frac{V_1 - V_2}{V_2}\right)$$

R_m is equal to the ohms-per-volt for the meter, multiplied by the full-scale reading on the range used. The ohms-per-volt is usually specified, but before using it for measuring purposes one should make sure that it is not merely a nominal figure.

A mains-frequency bridge of the type described in Sec. 7.14 covers a wider range than most ohmmeters and is nearly as convenient, but cannot be used if the thing being measured is reactive, i.e., has much capacitance or inductance. If it has, no clear balance is possible, unless of course it is being compared with a standard component of the same kind connected to the " Match " terminals. And it must be remembered that some resistances are distinctly different from the z.f. values, even at as low a frequency as 50 c/s. Another limitation of a bridge is that it cannot be used when the quantity being measured is drifting or fluctuating.

9.2. THE WHEATSTONE BRIDGE

The most suitable instrument for accurate measurement of z.f. resistance—provided that it is not varying—is the Wheatstone bridge (Sec. 7.2). Fig. 9.2 shows a practical form, either as a permanent self-contained combination, or as one of the adaptations of a general-purpose bridge (Sec. 7.4), or made up as required. If temporarily connected—especially when measuring low resistance—care must be taken to reduce the resistances of connections to a minimum.

In principle the measurement is simple: R_x is connected to the appropriate terminals, and P, Q and R are adjusted until the galvano-

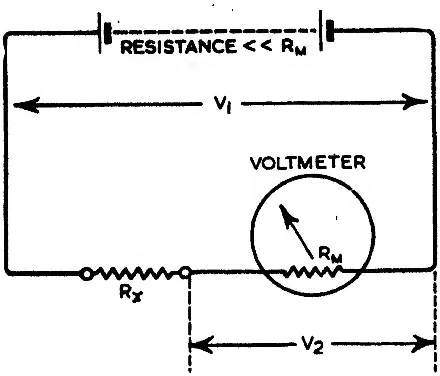


Fig. 9.1—Measurement of resistance by voltmeter

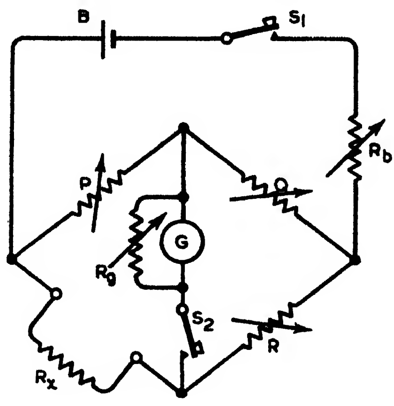


Fig. 9.2—Practical circuit of Wheatstone bridge for measuring z.f. resistance

meter G indicates balance (no current); then $R_x = RP/Q$. Until balance is approached, it is essential for the well-being of G to have R_g small or R_b large, or both. S_1 is closed first, to give time for any " kick " due to reactance in the R_x arm to subside before S_2 is closed. As the reading on G is reduced by arm adjustment, R_g (and perhaps the voltage of B) can be increased and R_b reduced to increase the sensitiveness of the indication, but of course R_b must never be reduced to the point where B would pass enough current to overheat any of the arms. The last drop of sensitiveness can be squeezed out by tapping S_2 in time with the natural swing of G; in this way a smaller current can be detected than if it is kept steady. It need hardly be mentioned that immediately balance is achieved, or whenever it is necessary to switch to a different ratio, the bridge should be restored to an insensitive condition.

Although G is not primarily intended as a measuring instrument, but merely to show the presence or absence of current, it can be used for interpolation to make up for the absence of a continuously variable resistance standard. Having found, for example, that with $R = 127\,\Omega$

the galvanometer reads 3 divisions to the left and with 128 Ω it reads 2 to the right, one can call the resistance 127·6 Ω.

Assuming the ratio arms P and Q are as described in Sec. 7.3, each switchable to 1, 10, 100, 1,000, and 10,000 Ω, and R is a four-dial decade box (Sec. 6.3), variable up to 11,110 Ω in steps of 1 Ω, one appears to have a bridge for measuring from about 0·0001 Ω to 100 M Ω. But the decade box does not provide adjustment to 1% below about 0·01 Ω, and apart from that it will be found in practice that the extreme values are not measurable, or at least not accurately, because of connection resistances and insufficient detector sensitiveness. If P and Q are in the ratio 10,000 : 1 the minimum observable deflection requires a much larger percentage unbalance than with a 1 : 1 ratio. Moreover, the ratio itself is not likely to be known to great accuracy when it is large, because low values of resistance are more affected by connection resistances.

9.3. AVOIDING ERROR IN BRIDGE MEASUREMENTS

Errors, especially those due to inferior equipment, can be largely avoided by using the substitution method (Sec. 3.8), the difference method (Sec. 3.9), or preferably both. With substitution, R_x and R are inserted in turn into the same arm. First R_x is balanced by adjusting a suitable precisely-variable resistance forming what is normally the R arm—or by adjusting the ratio, if that is variable. R is then substituted for R_x and adjusted until balance is restored; whereupon $R_x = R$. It is not necessary to know anything except R, so that is the only part of the bridge that need be good; even the ratio makes no difference except in so far as it affects the sensitivity of indication and the balancing resistance that can be used. But of course the range of measurement is restricted to that of the standard, R. In the difference method both R and R_x are kept in circuit all the time. Any increase (or decrease) in R_x is balanced by a decrease (or increase) in R. If both are in the same arm, this is just a variety of substitution, having the advantage that any incidental resistances such as leads are the same in both balances and so are eliminated by taking the difference between the " before and after " values of R. If they are in different arms this difference has to be multiplied by the bridge ratio, so the field of error is greater, but so also is the range of measurement. The advantages of the difference method may not be so very conspicuous when measuring z.f. resistance; they are much more so in r.f. measurements of all kinds, where there is otherwise great difficulty in excluding undesired influences.

9.4. LOW RESISTANCE

The ordinary form of Wheatstone bridge is not suitable for measuring very low resistances with accuracy, for which special forms are described in the large books, such as F. K. Harris's *Electrical Measurements* (Chapman & Hall).

The multi-range ohmmeter shown in Fig. 5.8 and described in

Sec. 5.8 is convenient for the low values for which the ordinary test set does not cater; but just as a voltmeter can be used for measuring the higher resistances a milliammeter or ammeter can be used in an analogous manner for low resistances, as shown in Fig. 9.3. The differences are that R_x is connected in *parallel*, and that the resistance

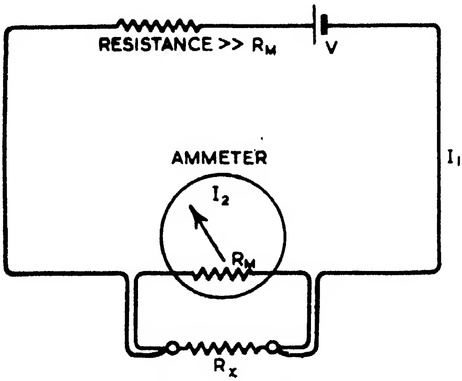


Fig. 9.3—Measurement of low resistance by shunted-meter method. Note arrangement of connecting leads to avoid displacement of zero

of the supply must be *large* compared with that of the meter so that the amount of current is not materially altered by inserting the meter. Then, if I_1 and I_2 are the readings before and after connecting R_x,

$$R_x = R_m \left(\frac{I_2}{I_1 - I_2}\right)$$

As with the voltmeter method, R_x is preferably not many times greater or less than R_m. Infinite R_x is indicated at the initial deflection, whatever R_m, and zero at zero deflection, provided that the resistance of the leads is eliminated by using *twin* flex joined only at the R_x end, as in Fig. 9.3. So the method is particularly good for testing uncertain resistances such as switch contacts, because the pointer can never be driven off the scale. And since the test is made at low voltage it shows up contact resistance that might be broken down by a higher potential. R_m (if not in the instrument's specification) can be found by connecting a known low resistance. For comparing switches it is not essential to know R_m.

9.5. VERY HIGH RESISTANCE

A convenient instrument for testing insulation resistance, especially for " field " use, is the Megger or one of its variants, described in Sec. 5.8 and needing no further reference.

Essentially the same method as that in Fig. 9.1 can be adapted for very high resistances by substituting for the voltmeter a sensitive galvanometer or microammeter in series with a source of known voltage and sufficient resistance to limit the deflection to full-scale when the R_x terminals are shorted. As before, only the *ratio* of meter readings need be known, but if the meter is calibrated in current it can first be used to measure the series resistance, if that is not known. E.g., a meter reading 1 μA full-scale from a 250-V source is thereby shown to have 250 MΩ in series. This is preferably a high-stability resistor, and above all must be free from risk of short-circuit. The effective range of this combination would be at least 25 to 2,500 MΩ, and shorting the R_x terminals would do no harm.

As an alternative to the delicate meter, or for carrying the range of measurement higher still, valve instruments are used. If a valve voltmeter with a high-resistance input is available, it can be adapted as a valve ohmmeter, along the lines explained in Sec. 5.18. In most valve ohmmeters R_x is connected in series with a standard resistance (usually one of several, providing as many ranges), with a valve voltmeter across one or the other. Some are similar to the ohmmeters previously described, in that the pointer remains on the scale for all values of R_x, but there are others that can be driven violently off the scale by connecting too low an R_x.

Full circuit details of the English Electric Co.'s instrument for testing insulation up to 10 kV are given by L. R. Hulls and K. A. Mackenzie (*E.E.*, Nov. 1952).

Instruments for measuring very high resistance can be checked by the method described at the end of the next section.

Bridges are not very often used for resistances of the many-megohm order, for the accuracy associated with them is difficult to achieve in these ranges, and is rarely required. But bridges described by A. B. Boff (*E.E.*, July 1950) and G. France (*E.E.*, Jan. 1957) measure resistances up to 100 million megohms. They use special electrometer valve voltmeters, and of course every precaution against leakage (Sec. 9.7).

An important thing to remember in high-resistance measurement is that most insulators and semiconductors fail to obey Ohm's law; their resistance depends on the voltage applied. Unless there are reasons for a different choice, 500 V should be used, and the reading taken after it has been applied for 1 minute, these being British Standard conditions. Their resistance also depends very much on temperature and humidity. Information on the properties of resistors up to 10^{13} Ω is given by G. France in the above reference. Measuring resistances as low as 100 MΩ is liable to considerable error, as shown by J. K. Wood in comparative tests (*E.E.*, June 1958).

9.6. CAPACITOR LEAKAGE

One of the commonest kinds of high resistance to be measured is the leakage resistance of capacitors. For this purpose a steady voltage is essential, as otherwise the indication is obscured by charging and discharging currents. And with low-leakage specimens much time is saved if there is provision for charging them to the measurement voltage through a reasonably low resistance.

Very high leakage resistance can be measured with simple apparatus by timing self-discharge through it from one voltage to another. The discharge formula (Sec. 14.6) is

$$RC = \frac{t}{2{\cdot}3 \log_{10} (V_1/V_2)}$$

where t is the time in seconds for C microfarads to discharge through R megohms from V_1 to V_2. If C is known, R follows; but actually RC is the real measure of the capacitor's goodness.

The apparatus needed is chiefly a valve with a very high input resistance, especially from grid to anode. If this is available as a d.v. valve voltmeter (Secs. 5.16–17), so much the better, as it will be already calibrated in volts; but since extreme absence of leakage in the grid circuit is more important than accuracy and stability of calibration it will probably be better to rig up a specially selected valve (Sec. 5.18) as in Fig. 9.4. The anode voltage should be kept down to about 50 V,

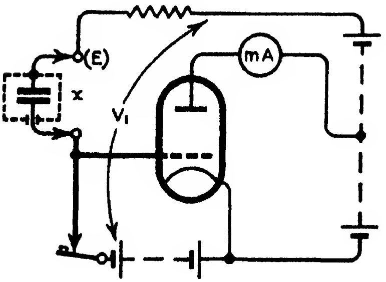


Fig. 9.4—Valve circuit for measuring capacitor leakage and other high resistance

or even less, to minimize grid current due to ionization, and as this is hardly enough for the leakage test voltage a separate h.t. connection is shown; the parts drawn in heavy line must of course be very highly insulated, being supported by a material such as polythene and having as little contact as possible with that. The grid bias is chosen so that when the switch is closed the pointer is not much above zero. The capacitor under test is charged by both anode and grid voltages (total V_1) through the safety resistance, which may be about 50 kΩ. When the switch is opened the capacitor discharges through its own leakage, the grid potential rises, and so does the anode current. The time taken for the pointer to reach a second mark, at which the capacitor voltage is V_2, is noted, and when divided by $2{\cdot}3 \log_{10} (V_1/V_2)$ gives RC in megohm-microfarads. The difference in grid voltage between the two marks ($V_1 - V_2$) should be chosen so that this multiplier is a convenient whole number. $V_1 - V_2$ should also be a small fraction of the initial voltage V_1, not only to save time but to avoid complications due to the voltage—and so possibly the resistance—varying substantially during the measurement. It is also necessary that it be small enough not to reduce the negative bias below at least 2 V, which is about where grid current begins to increase rapidly. In Table 9.1 the fraction $(V_1-V_2)/V_1$ is worked out for a few suitable values of the factor RC/t by which t has to be multiplied to give RC. E.g., if the initial charging voltage V_1 is 200 V, 10 of this being negative grid bias,

Table 9.1

$\frac{RC}{t}$	$\frac{V_1 - V_2}{V_1}$
500	0·0020
200	0·0050
100	0·0100
50	0·0198
20	0·488

and the factor 100 is chosen, the second mark on the meter scale must correspond to the anode current when the bias is reduced by 2 V. Then, if when testing a capacitor the time taken for the pointer to move from one mark to the other is 6 seconds, the MΩ-μF figure is $6\times100=600$; and if the capacitance is 0·1 μF its leakage is 6,000 MΩ.

For most purposes it is enough to see whether the pointer flicks rapidly across the scale, moves at a pace that is easily followed, or crawls almost imperceptibly, corresponding to bad, average, and exceptionally good capacitors respectively. Some idea of the leakage of the apparatus can be obtained by opening the switch very momentarily with nothing connected. If the pointer is at all sluggish in getting off the mark, the capacitance of valve electrodes and connections being of the order of 10 pF, leakage can be considered satisfactory.

Non-capacitive high resistance can be measured by connecting it in parallel with as perfect a capacitor as possible and comparing the resistance of this combination (say R_1) with the resistance of the capacitor alone (R_2). If R_x denotes the resistance to be found,

$$R_x = \frac{R_1R_2}{R_2 - R_1}$$

It should be realized that the measured leakage of most capacitors varies for some time after applying a charging voltage, so the test should be repeated until consistent results are obtained. There is also a " soaking " effect, due to redistribution of charge within a capacitor, which reduces the accuracy of discharge methods.

9.7. GUARD-RING TECHNIQUE

When measuring very high resistance, it is necessary to make sure that the result is not going to be affected by other leakages. Fig. 9.5

represents a typical apparatus, with its source of test voltage, and an indicator, which in this case is shown as a valve voltmeter reading the drop caused by leakage current through a known resistance R_S, but it might be a sensitive galvanometer. R_X is the resistance to be measured, connected between the test terminals 1 and 2. R_1 represents unavoidable internal leakage, and R_E is external leakage. Without special precautions, the indicator would read current through these leakages as well as through R_X. But if terminal 2 is surrounded by a

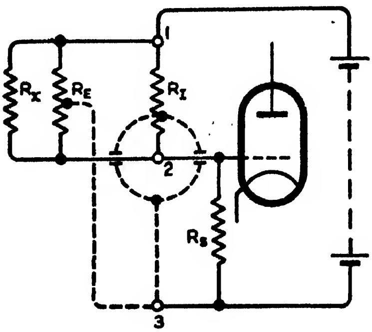


Fig. 9.5—The dotted lines show guard arrangements to intercept leakage by paths it is not desired to measure. R_I represents internal leakage and R_E external leakage

metal ring connected to 3, current from 1 passes straight to 3 and is not indicated. Similarly current through R_E can be intercepted by an externally connected guard ring. One must take care, however, that the leakage resistances between the guard rings and terminal 2 are not low enough to shunt R_S appreciably. This is hardly likely to happen if R_S is the resistance of a galvanometer, but care may be necessary with a high-resistance valve voltmeter, which can give very misleading results from this cause.

For a fuller exposition, see " Measuring High Resistance " (*W.W.*, June 1952) and the references given in Sec. 9.5. A good example of guard-ring technique is the English Electric insulation tester mentioned in Sec. 9.5.

9.8. RESISTIVITY OF LIQUIDS

The resistivity, ρ, of a material can easily be derived from the resistance of a sample of known and uniform length (l) and cross-sectional area (A):

$$\rho = R\frac{A}{l}$$

When the material is a liquid, two complications arise: (1) avoiding polarization effects, and (2) making contact with a suitable sample. The first is easily solved by using a.c., preferably not very low in frequency; e.g., 1,000 c/s rather than 50 c/s. The second depends on the order of resistivity; if low, as for example battery acid or mercury, an appropriate arrangement would be a glass tube of uniform bore, with electrodes of some metal that is not attacked by the liquid. With liquids of higher resistivity, to bring the resistance to be measured within a favourable range it may be necessary to use relatively large-area plate electrodes close together. The problem of accurately determining the effective length and cross-sectional area of such a sample can be neatly dodged by making use of the fact that the ratio A/l is equal to C/ϵ, where C is the capacitance between the same electrodes in a medium of absolute permittivity ϵ. By measuring C with the plates in air, for which ϵ is very nearly $1/(36\pi \times 10^9)$, ρ can be found even though A and l are non-uniform and subject to "edge effect":

$$\rho = 36\pi RC \times 10^9 \text{ ohm-metre} \qquad [\Omega\,;\,\text{F}]$$

It is necessary, of course, for the electrodes to be rigidly mounted on one or more insulators, and in fact the system may conveniently be a small air capacitor. But when measuring R the insulation contributes negligible conductance, whereas during the measurement of C it contributes more than its share of capacitance. This cause of error can be substantially removed by using a variable capacitor, measuring capacitance and conductance at, say, maximum and minimum settings and taking the differences. Conductance is the relevant quantity, because it is the analogue of capacitance, but most instruments are calibrated in resistance, from which conductance is derived by taking the reciprocal. So if R_1 and C_1 are readings taken at maximum interleaving of the plates, and R_2 and C_2 at minimum,

$$\rho = 36\pi \times 10^9 (C_1 - C_2)\Big/\left(\frac{1}{R_1} - \frac{1}{R_2}\right) \text{ ohm-metre} \qquad [\Omega\,;\,\text{F}]$$

$$\left.\begin{aligned} &= \frac{0{\cdot}113(C_1 - C_2)R_1R_2}{R_2 - R_1} \text{ ohm-metre} \\ &= \frac{11{\cdot}3(C_1 - C_2)R_1R_2}{R_2 - R_1} \text{ ohm-cm.} \end{aligned}\right\} \qquad [\Omega\,;\,\text{pF}]$$

The plates must obviously be of a metal not attacked by the liquid, and if any of the insulation is immersed its properties must be considered too; e.g., polystyrene would be unsuitable in a liquid plasticizer.

9.9. CAPACITANCE: MEASUREMENT BY VOLTMETER

The voltmeter method shown in Fig. 9.1 for measuring resistance can be used for capacitance if an a.c. source and meter are substituted for the d.c. types. An ordinary rectifier type of a.c. voltmeter is suitable, if used on one of the higher ranges where its multiplier resistance is large enough to swamp the uncertain rectifier resistance; and the mains are a suitable source. The measuring procedure is the same as for resistance, V_1 being the voltmeter reading direct from the source, and V_2 with C_x in series. The calculation has to allow for the phase difference between the voltages across C_x and across the meter (resistance R_m), and as C_x is assumed to be pure reactance, X_c, this is taken as 90°·

$$X_c = R_m\sqrt{\left(\frac{V_1}{V_2}\right)^2 - 1}$$

from which $$C_x = \frac{159\,V_2}{fR_m\sqrt{V_1^2 - V_2^2}} \qquad [\text{c/s; k}\Omega\text{; }\mu\text{F}]$$

If f is always the same, say 50 c/s, this method can be used to provide a multi-range test set with a scale reading capacitance directly. With a typical meter, having about 100 kΩ resistance on a 250-V range, the best part of the capacitance range is about 0·002 to 0·1 μF (2 to 100 nF).

9.10. CATHODE-RAY-TUBE METHOD

The voltages across C_x and a known resistance can be compared by a cathode-ray tube, and the advantage of this over using the voltmeter as just described is that it shows not only the impedance of the capacitor but also its phase angle. So besides being applicable to ordinary

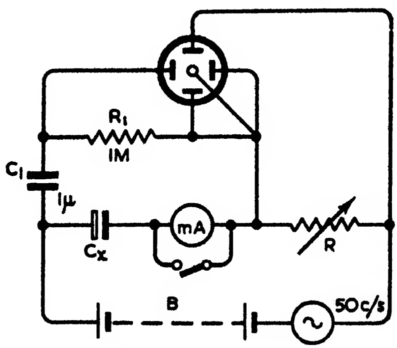


Fig. 9.6—Method of measuring capacitance, phase angle, and leakage of electrolytic capacitors by cathode-ray tube, with polarizing voltage applied

mica and paper types which normally have a phase angle of almost 90°—power factor or dissipation factor or loss tangent (Sec. 6.2) nearly zero—it is useful for electrolytic capacitors whose phase angle may be substantially less than 90°, which rules them out from the voltmeter method. It is also easier to apply a polarizing voltage, and Fig. 9.6 shows the c.r.t. method with the elaborations needed for this: B to supply a polarizing voltage exceeding the peak a.v. across C_x; a milliammeter to measure the leakage current, with switch for short-circuiting it to prevent damage by the surge when B is first applied; and C_1 and R_1 to block the polarizing voltage from the c.r.t. For non-electrolytics, all these can be omitted. In either case, R is adjusted until the deflection due to it is equal to that due to C_x. Final adjustment of R can be done more precisely by observing the deflections one at a time, the plate not in use being connected to anode. Assuming the a.c. source is free from harmonics, a perfect capacitor will show a perfect circle on the screen, and as resistance and reactance are then equal

$$\begin{aligned} C_x &= 1/\omega R \\ &= 159/fR \qquad \text{[c/s; k}\Omega\text{; }\mu\text{F]} \end{aligned}$$

Unequal deflection sensitivities of X and Y plates can be allowed for by taking a second reading with the X and Y connections changed over. With R_1 and R_2 denoting these two readings, $R = \sqrt{R_1R_2}$.

Assuming the frequency is 50 c/s, and neglecting the series resistance of the capacitor, Table 9.2 gives a few representative values. For smaller values of C_x, it is better to use a higher frequency. And for the largest values it may be necessary to amplify the deflections, so as not to exceed the voltage ratings; and then amplifier phase shift must be considered.

Table 9.2

C_x (μF)	R (kΩ)
0·1	31·8
1	3·18
8	0·398
25	0·127
100	0·0318

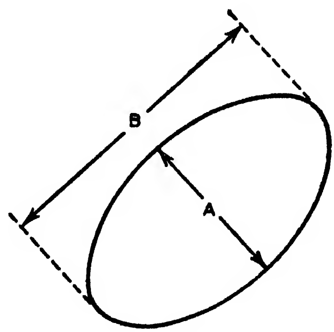


Fig. 9.7—Dimensions of c.r.t. ellipse from which phase angle can be calculated

If the phase angle ϕ between the voltages across R and C_x is less than 90° the figure is a diagonal ellipse, as shown in Fig. 9.7, and ϕ can be found from the relationship

$$\tan\frac{\phi}{2} = \frac{A}{B} \qquad (W.W., \text{Oct. 1952, p. 432})$$

A and B being the minor and major axes respectively. It is absolutely necessary for the X and Y deflections to be equal, but for this purpose the relative deflection sensitivity is unimportant. If, as is usual, ϕ is nearly 90°, it is better to measure the angle δ ($= 90° - \phi$) directly by substituting for R a variable capacitor having negligible loss angle; the formula is the same except for δ taking the place of ϕ. This method is better also for measuring C_x, but was not prescribed above because it was taken for granted that a continuously-variable negligible-loss capacitor up to many μF would not be available. If it is, of course one would not use R. There is an alternative method of arriving at δ which does not necessitate equality of deflection, so any negligible-loss capacitor of the same order of capacitance can be used for the comparison. The phase angle (δ) between the two capacitors is calculated from the dimensions shown in Fig. 9.8:

$$\sin\delta = C/D$$

C and D are difficult to measure when only a single ellipse is produced on the tube screen, but G. N. Patchett has shown (*E.E.*, April 1944) how this can be overcome by using a double-beam c.r. tube, with the two Y plates connected together. The result is the double ellipse,

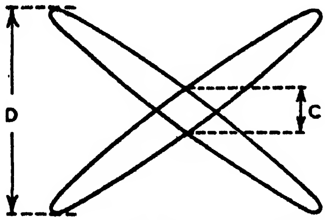


Fig. 9.8—Double-beam c.r.t. trace, more convenient than Fig. 9.7 for measuring

enabling C and D to be measured straight off the tube with great ease.

Capacitance and phase angle can also be measured using a valve voltmeter instead of a c.r. tube, as described in Sec. 9.19.

9.11. BRIDGE METHODS FOR CAPACITANCE

The three most important basic types of bridge for measuring capacitance are shown in Fig. 7.13: *b*, *c*, and *d*. Types *b* and *c* can be understood to include those in which P and Q are replaced by transformer ratio arms (Sec. 7.10). (In addition type *h*, being reversible, can be used to measure capacitance as negative inductance; see end of Sec. 9.22.) No capacitor is pure capacitance, so the one being measured is shown as a dotted ring containing resistance r_x as well as C_x. Actually the losses represented by r_x are distributed, and it is for the sake of uniformity that in every case it is shown in series. This is the most convenient representation in types *b* and *d*, but in *c* it results in equations that are awkward* so is not used. If the unknown is represented in *c* by C_x with parallel resistance R_x, the equations for these are the same simple ones as for *b*. Actually it is the ratio of resistance to reactance, or loss tangent (tan δ),† that is the more useful

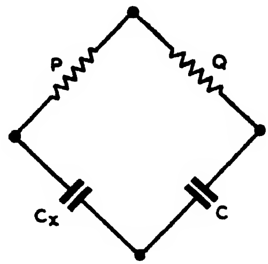
* For measuring capacitance, but not for frequency; see Sec. 7.19.

† See end of Sec. 6.2. Loss tangent is the same thing as dissipation factor but briefer, and practically the same thing as power factor when it is small. δ is the loss angle, $= 90° - \phi$, where ϕ is the phase angle between voltage and current.

expression, and this remains as shown in the equations in Fig. 7.13 regardless of how capacitance (or inductance) and resistance are mixed in the dotted circles.

In capacitors that are any good at all (except electrolytics), tan δ is less than 0·01, so the less refined capacitance bridges hopefully ignore it. If this is done, types *b* to *d* all reduce to the same thing (Fig. 9.9). The mains-frequency bridge described in Sec. 7.14 is an example of

Fig. 9.9—If capacitor resistance can be neglected, the three bridges in Fig. 7.13 b-d *all reduce to this*



this, on its lower ranges. But the 1 μF range, being used for measuring electrolytics with their comparatively high series resistance, of necessity includes a resistance balance (type *b*). Instructions for using this bridge are included in Sec. 7.14. Although capable (if carefully made) of checking components on its 100-pF range, it is not at its best for small capacitances.

For precise measurements it is necessary to allow for resistance, even in " good " capacitors, if only to get a sharp balance. Sometimes tan δ is of greater interest even than the capacitance.

9.12. THE pF RANGE

If a permanent set-up is required for precise measurements, with emphasis on tan δ, the Schering bridge (Fig. 7.13*d*) is the usual choice. C_r can be directly calibrated in tan δ (for a given frequency), and C_x can be balanced either by fixed P and variable C or vice versa, the latter being especially convenient for high-voltage tests. For accuracy, however, it is better to have equal fixed-ratio arms; then the zero-reading capacitance of C_r can be balanced by a small pre-set capacitance across P.

The general-purpose bridge of Sec. 7.4 (Fig. 7.3) can be used to make type *b* by connecting a standard capacitor to the S terminals to act as *C*, and with the switch toward *x* the resistance box serves as *r*. To avoid complications, the standard should be good enough for its losses to be negligible, and if the ratio arms are switched in steps it must be continuously variable; both conditions call for an air capacitor, which is usually limited in range to about 1,000 pF. For measuring capacitances smaller than this, the ratio can be fixed at 1 : 1, which simplifies guarding against stray admittances (Sec. 7.8). Even so, if the normal bridge use is adopted, fairly comprehensive screening such as Fig. 7.6 is necessary for accurate work. But difference substitution is even more advantageous than with resistance (Sec. 9.3), for in conjunction with a simple Wagner earth (Sec. 7.7) it removes the necessity for elaborate screening. Fig. 9.10 shows the whole arrangement. C_B is any ordinary variable capacitor of about the same maximum as C. First the switch is set to position 1 so that the Wagner earth tapping T can be adjusted to exact centre, or at any rate to match P and Q. If there is any difference between the stray capacitances across the two halves of the bridge, a small amount of capacitance across one half or the other of the Wagner earth resistance will be needed for exact balance, but for ordinary purposes such a refinement is hardly necessary. On the assumption that the standard, C, has less resistance than C_B, the switch is then moved to 2, and, with C_x disconnected as shown, C is set at or near maximum (reading C_1) and balanced by adjusting C_B and r (reading r_1). C_x is then connected, and C and r readjusted to balance (readings C_2 and r_2), switching over to 3 if necessary. It may be as well to check Wagner again before finally adjusting. Denote $C_1 - C_2$ by ΔC, and $r_1 - r_2$ (or $r_1 + r_2$ if the switch has had to be moved) by Δr. The exact formulae for the results are rather complicated, but provided that the square of tan δ is negligible compared with 1 (which is usually the case) they simplify to:

$$C_x \simeq \Delta C$$
$$r_x \simeq \Delta r\,(C_1/\Delta C)^2$$
$$\therefore \tan\delta \simeq 2\pi f \Delta r\, C_1^2/\Delta C$$

If f is always 1 kc/s and C_1 is 1,260 pF, and Δr is in ohms and ΔC in pF, tan δ further simplifies to $\Delta r/\Delta C\%$. This method has the great advantage that defects in all other arms are eliminated, and so is even the imperfection of the standard capacitor, if, as is justified with any reasonably good type, it can be represented at a given frequency by a constant and very high resistance in parallel.

The main possible cause of error to be guarded against is the capacitance of the connecting leads. This problem is fully discussed in a valuable article in *G.R.E.*, July 1959. It refers to the method by R. F. Field in the May 1947 issue, which is suitable for general purposes where an uncertainty of the order of 0·2 pF can be tolerated. According to it, while the first reading is being taken C_x ought to be in position as in Fig. 9.10, connected at the earth end, and with a thin but rigid lead ready attached to the other end of C, and well spaced from everything except the corresponding C_x terminal, which it should approach to within about ¼ in. For the second reading C_x is connected by closing this gap. The capacitance of this lead then affects both readings of C almost exactly equally, so is eliminated in the difference. When measuring a two-terminal capacitor, some slight uncertainty in its capacitance is inevitable, so for precision better than 0·2 pF it is necessary to use a three-terminal capacitor, as in Fig. 9.11. Its capacitance is defined as the direct capacitance between the terminals,

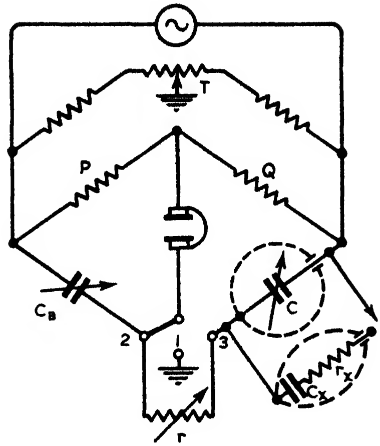
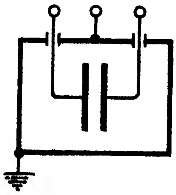


Fig. 9.10 (right)—Arrangement of bridge for difference-substitution method of measuring capacitance and loss angle up to about 1,000pF

Fig. 9.11 (below)—Three-terminal capacitor, which must be used in measurement circuits in which capacitances to earth are rejected



and is measured with a transformer bridge or any other in which the capacitances from terminals to screen are excluded from the result; see Sec. 9.16.

9.13. TESTING DIELECTRIC MATERIALS

The methods just described are suitable for measuring the loss tangent (tan δ) and dielectric constant or specific permittivity (ϵ_s) of materials. The test is made on a sample in the form of a sheet of uniform and preferably fairly small thickness—say not more than about 0·1 in—which is made into a capacitor by means of an electrode on each side. Perhaps the most convenient method (if wetting the sample is allowed) is to paint colloidal graphite disks on to the sheet and back them up with brass or copper plates. An alternative to the graphite is tinfoil stuck on with the least trace of Vaseline and pressed into close contact. It is important to have perfect contact with the sample, so that the measured tan δ and ϵ_s are the true properties of the material. Detailed procedure for ensuring high accuracy is given in B.S. 903 and in Hartshorn's *R.F. Measurements*, p. 192. The formula for ϵ_s is

$$\epsilon_s = 11{\cdot}3\, C_x\, t/A$$

where C_x is the capacitance of the sample in pF, t the dielectric thickness in cm and A the effective area of dielectric between the electrodes in sq cm, $= \pi D^2/4$ if it is circular with diameter D.

A complication is that the effective area A is greater than the area of the electrodes, because of the stray field around the edges. Elaborate formulae have been produced for calculating this, but one must make sure that the conditions which have been assumed apply in one's own work. The Kirchhoff formula quoted by Hartshorn implies that the dielectric fills all the space around the electrodes, so applies only to measurements on liquids. The formula is then unnecessary, because ϵ_s is equal to the ratio of capacitance with liquid to that with air (which in this case can be measured, as the electrodes must be fixed). Other formulae assume that the dielectric (solid) ends at the edges of the electrodes, so that the whole of the extra field is in air. In practice it may be inconvenient to cut the sample exactly to this size, and a third condition, not readily calculable, is with the sample extending at least several times its own thickness all round beyond the electrodes. Provided that the diameter is more than about 30 times the thickness, the edge correction is quite small, and the error in approximating to

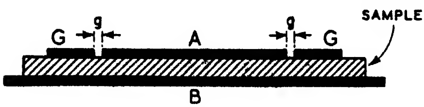


Fig. 9.12—Section of electrode system for measuring permittivity of a sheet sample of solid material

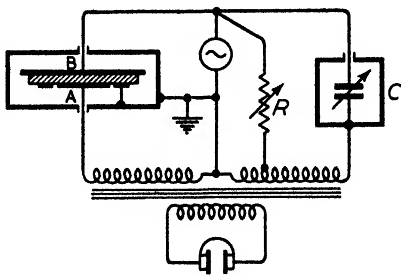


Fig. 9.13—Application of transformer ratio-arm bridge to electrode system shown in Fig. 9.12 for measuring permittivity

it by defining D as the actual diameter of the electrodes plus t is likely to be less than that of measuring t, especially if it is not perfectly uniform.

If the ratio D/t is unavoidably less, a guard ring should be used; G in Fig. 9.12. There is still an edge correction, but provided the gap g between A and G is small it is sufficient for all ordinary work to take D as the actual diameter of A, plus g. The capacitance must be measured between A and B, with G at the same potential as A but not included in the measurement. The Blumlein bridge is particularly convenient for this, and Fig. 9.13 shows one scheme. At balance there is no detector voltage, so A and G are at the same potential, as required. For low-loss materials the balancing resistance R would be very large if in parallel with C, so it is reduced by tapping down as shown. Alternatively a small resistance can be used in series with C.

A rather elaborate set-up for precise measurements of permittivity is described by A. M. Thompson in *Proc. I.E.E.*, Part B, Nov. 1956.

Another method of allowing for edge capacitance, suggested by G. W. Short, is based on the assumption that it is proportional to peripheral length, whereas the capacitance directly between the electrodes is of course proportional to the square of their diameter. So a first capacitance reading, C_1, is taken with the largest practicable electrodes that leave a several-times-thickness margin of dielectric sample all round. A second reading, C_2, is then taken with electrodes one nth the area. The true capacitance with the large electrodes (i.e., $C_1 - C_{edge}$) is then $n(C_1 - nC_2)/(1 - n)$. If the small electrodes have half the area of the large ($n = 2$), this simplifies to $2C_1 - 4C_2$. Most of the edge field being in the dielectric, tan δ can be regarded as that measured in conjunction with either C_1 or C_2; comparison of these two values, which should be equal, gives a check on the work.

Although tan δ varies much less with frequency than r_X, and tests made at a.f. usually give some idea of how the same materials compare at r.f., it is best to test at the working frequency. Very small capacitance is easier to measure at r.f. For r.f. methods see Secs. 9.25 to 9.35.

9.14. LARGER CAPACITANCES

As a continuously variable capacitance standard is not usually available much above 1 nF, the 1:1 ratio bridge methods in Sec. 9.12 may not be practicable for the higher ranges, and it is necessary either to use the variable air standard with higher ratios or a fixed standard with continuously variable ratio (e.g., Fig. 7.14). The disadvantage of the first is that whereas the *difference* in capacitance between two settings of a variable standard can be known very accurately, the actual capacitance at any setting is uncertain when it is connected in a bridge, because an unknown amount of stray capacitance is added, and this is multiplied on the higher ranges by the bridge ratio. So it is usually better to have a relatively large fixed standard capacitance so that strays are negligible—preferably a capacitance of the same order as C_X—and to balance with a variable ratio arm. This is satisfactory so long as tan δ does not approach 0·5.

One method of measuring a capacitance that is above the range of the bridge is to connect it in series with some capacitance that is within the range. If C_1 is this capacitance by itself, and C_2 when it has C_X in series, then

$$C_X = \frac{C_1C_2}{C_1 - C_2}$$

Apart from errors due to stray capacitances (which can be minimized by connecting the relatively large C_X on the low-potential side of C_1) it must be realized that errors in the readings are likely to be magnified in the result.

9.15. ELECTROLYTIC CAPACITORS

Fig. 9.14 shows modifications for applying a polarizing voltage to an electrolytic capacitor while it is being measured. Incidentally the standard, C, must be capable of withstanding the same voltage. Phones, if used, had better be kept from the polarizing voltage by transformer coupling. If a milliammeter is used for measuring leakage it must be

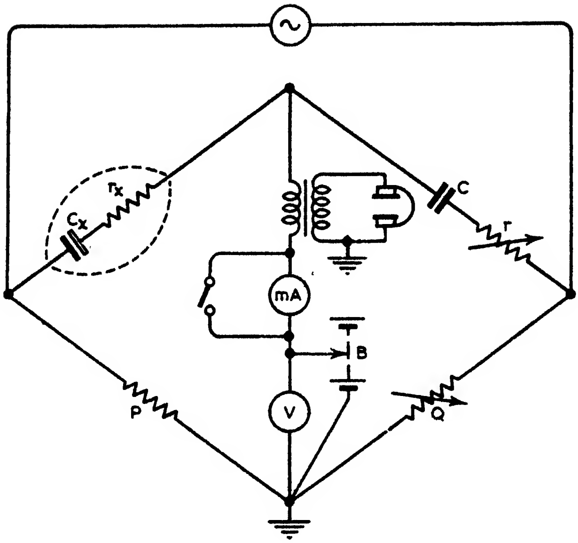


Fig. 9.14—Modifications of capacitance bridge to allow a known polarizing voltage to be applied to the capacitor under test

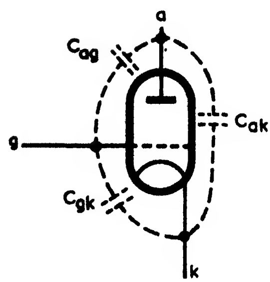


Fig. 9.15—Nomenclature of the three interelectrode capacitances of a triode

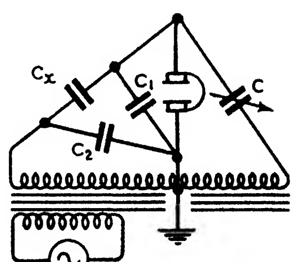


Fig. 9.16—Use of a transformer ratio-arm bridge for measuring direct capacitance (C_x) of a component that has capacitances (C_1 and C_2) or other admittances to earth

short-circuited during charging. In other respects operation is the same as for the Sec. 7.14 bridge. See also Sec. 9.10.

9.16. DIRECT CAPACITANCE

Sometimes the points between which capacitance is to be measured have capacitance to some third party or parties; for example, valve electrodes, or wires in a cable with a conducting sheath. Consider a triode, where there are three interelectrode capacitances, shown diagrammatically in Fig. 9.15. It might be supposed that all one has to do is to connect the two electrodes concerned to the measuring apparatus and leave the other unconnected. But this would not give the correct result. One procedure is to measure in turn the capacitance from each electrode to the other two joined together. For instance, grid and anode are tied together and the capacitance C_1 from them to cathode is measured:

$$C_1 = c_{gk} + c_{ak}$$

Similarly $$C_2 = c_{ag} + c_{ak}$$

$$C_3 = c_{gk} + c_{ag}$$

The required capacitances are found by solving these simultaneous equations, thus:

$$c_{gk} = (C_1 + C_3 - C_2)/2$$

$$c_{ag} = (C_2 + C_3 - C_1)/2$$

$$c_{ak} = (C_1 + C_2 - C_3)/2$$

These are called the *direct* capacitances between the electrodes.

With a suitable bridge, it is possible to eliminate undesired capacitances—or in fact any sort of admittances—and measure the direct capacitance at one go. The most effective bridges for this have transformer ratio arms (Sec. 7.10). In Fig. 9.16, C_x is the direct capacitance and C_1 and C_2 are capacitances to somewhere else. These should be tied to the earth point on the bridge. C_1 is thereby brought across the detector, which may reduce its sensitivity but cannot affect the balance; and C_2 comes across part of the transformer, which, being very closely coupled, is effectively across the source and again has no effect on balance. This is true if C_1 and C_2 are any kinds of admittance; some bridges on the market will tolerate admittances hundreds of times greater than that of C_x, so that quite small capacitances can be measured without disconnecting them from their circuit. But this demands very careful design, as explained in *Proc. I.E.E.*, Part III, May 1949, p. 189.

If, as often happens, one terminal of C_x is earthed, the earth shown in Fig. 9.16 should be moved to the other end of the detector, leaving any " third parties " connected as shown.

For an example of a guard-ring connection, see Fig. 9.13.

9.17. DIRECT-READING CAPACITANCE METERS

For rapid component checking, even the small amount of adjustment needed to get a reading on a bridge of the kind described in Sec. 7.14 becomes tedious, and the "ohmmeter" method (end of Sec. 9.9) is limited in range. For pF values, 50 c/s is really too low. One solution, adopted in the "Picomat", is to use a relatively high-frequency transistor oscillator, with a step-up transformer to give a direct reading on a crystal-rectifier meter (0 to 50 μA) in an "ohmmeter" type of circuit, using an all-capacitor potential divider. Two ranges cover 0 to 500 pF and 10 pF to 10nF. The power source is a small 6 V accumulator, with provision for trickle-charging it from the mains.

In Sec. 7.19 reference was made to direct-reading frequency meters working on the pulse-counting principle. The ranges are obtained by switching known capacitances. The same principle can be used to measure capacitance directly, the ranges being obtained by switching known frequencies.

Basically similar is the method used in the Heathkit CM-IU capacitance meter, which has four direct-reading scales reading from 0·1 μF down to about 2 pF. Square waves are generated by a cathode-coupled multivibrator, the frequency of which is controlled in four steps by the range switch. The capacitance to be measured is connected in series with the rectifier meter and a 10 kΩ resistor across the square-wave output to form a differentiating circuit. The pulse width, and hence the average meter current, is directly proportional to C_X.

The design and construction of a rather more specialized apparatus, for measuring circuit stray capacitances in the range 0 to 300 pF, is given by J. C. S. Richards (*E.E.*, March 1957). Advantages are that they can be measured in parallel with comparable resistance, and up to 3 ft of coaxial cable can be used for connecting to the unknown *in situ.* The heart of the instrument is a hexode valve, to the third grid of which an r.f. signal is applied directly. It is also applied across the test circuit to the first grid, and the principle is the rapid phase change (indicated on an anode-current meter) as the test circuit is swung through resonance.

9.18. INDUCTANCE: DIFFICULTIES OF MEASUREMENT

Measurement of inductance presents more difficulties and complications than capacitance. (For how these affect the production of standards, see Sec. 6.9.) Inductors to be measured can be roughly divided into two groups: air cored (and iron-dust or ferrite cored) coils of the microhenry order, mainly for r.f. tuning; and iron-cored coils of the henry order, for a.f. and power frequencies. Measurement of r.f. coils at r.f. is complicated by their self-capacitance; at a.f. this is negligible, but unfortunately the inductive reactance also is small, usually less than the resistance of the wire. At 1 kc/s an ordinary medium-frequency coil has a reactance of about 1 Ω, and at 50 c/s only 0·05 Ω. Television and other v.h.f. coils are worse still, and they have so few turns of wire that each half-inch of connecting lead makes a difference.

The chief difficulty with iron-cored coils is that an inductance figure means very little unless supplemented by details of the amplitude of a.c. used for measuring, the amount of d.c. flowing at the same time, and the method of measurement. The inductance varies not only with the amount of a.c. and d.c. but even during each cycle of a.c. This raises both theoretical and practical problems. What *is* the inductance, when it is varying all the time? The matter can be simplified, at least in theory, by passing through the coil a sine-wave current I, measuring the value V and phase difference ϕ of the *fundamental* of the resulting voltage across it, and reckoning the inductance as $V \sin \phi/\omega I$. In practice, unless there is relatively infinite linear impedance in series with the coil, the non-linearity of the inductance causes the current waveform to be distorted, and in general the current harmonics that accompany the fundamental alter the inductance. Moreover, unless a sharply tuned detector is used the presence of the harmonics makes it difficult or impossible to obtain a clear indication of balance.

9.19. THE THREE-VOLTAGES METHOD

The fact that practical inductors are not even approximately pure reactance rules out the simple voltmeter method used for capacitance (Sec. 9.9), in which a 90° phase angle could reasonably be assumed.

The c.r. tube method (Sec. 9.10) is inconvenient unless about 20 V or more can be set up across the coil under test, and the resulting iron-core distortion is usually sufficient to prevent the trace from being used as a basis for measurement. But a similar method using a valve voltmeter is more generally useful for iron-cored coils and even air-cored coils down to a few millihenries—and incidentally is also very suitable for measuring capacitors that cannot be assumed to have nearly 90° phase angle. See Fig. 9.17. R, a known non-reactive resistance, is preferably of the same order as the impedance of the coil, and must be capable of carrying any d.c. it may be desired to pass through the coil. When the a.c. and d.c. are adjusted, the voltmeter is connected in turn across R, across the coil, and across both. Call the readings V_1, V_2, and V_3 respectively. The quickest way of finding the result is to draw the impedance diagram, of which Fig. 9.18 is an example. Horizontal distance represents resistance and vertical distance reactance. R, being a pure resistance, can be represented to any convenient scale by the line AB. The impedances of the coil, with and without R in series, are known in magnitude but not phase. With centre B and radius RV_2/V_1, and with centre A and radius RV_3/V_1, draw arcs cutting at C. The vertical CD represents to the same scale the reactance of the coil, and DB its resistance, while ϕ is the angle of lag. $L_x = X_L/2\pi f$.

The advantage of making R about equal to the impedance of the coil is that then the valve-voltmeter calibration does not have to be

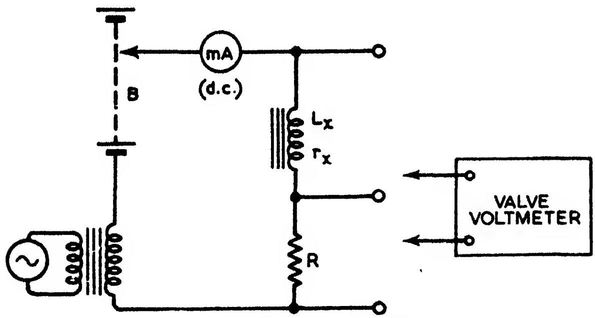


Fig. 9.17—The three-voltage method of measuring reactance and phase angle, shown here applied to an iron-cored inductor with superimposed d.c.

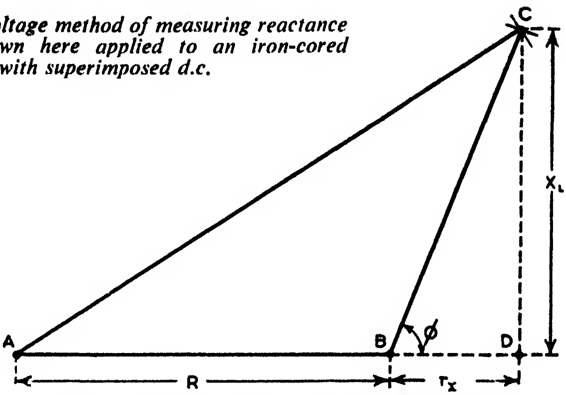


Fig. 9.18—Diagram used for deriving reactance (X_L) and phase (ϕ) from voltmeter readings in Fig. 9.17

accurate over a wide range. If it is, however, ϕ can be found more accurately by making R considerably greater. It should never be less. If an iron-cored coil is run under conditions that tend to distort the current waveform, the distortion can be reduced as much as one likes by making the resistance in series (not necessarily all in the position R) sufficiently large. One way of doing this is to put the coil and R in the anode circuit of a pentode. But then the voltage waveform of V_2 will be very distorted, so to give correct results it is necessary to put a sharply tuned filter, or at least some harmonic-reducing shunt capacitance, in front of the voltmeter. The effect on the calibration does not matter, as all readings will be affected in the same ratio. With this precaution the method fulfils the requirements stated in Sec. 9.18. Frequency should be low enough for the effect of self-capacitance to be negligible; 50 c/s is usually safe in this respect.

One is sometimes told that r_x can either be neglected altogether when measuring l.f. inductors, or assumed to be the same as the z.f. resistance, or not much more. Even at 50 c/s this is seldom true, r_x usually being many times greater than the z.f. value.

9.20. CAPACITANCE-COMPARISON METHOD

Another method of measuring inductance, due to H. M. Turner (*Proc. I.R.E.*, Nov. 1928), convenient if one has capacitance variable from 0·01 μF to 1 μF or more, and suitable for inductances from about 1 H upwards, is shown in Fig. 9.19, again with provision for

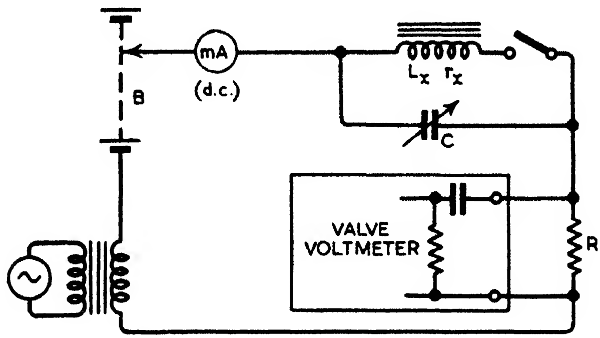


Fig. 9.19—H. M. Turner's method of measuring inductance, using large variable capacitance

superimposing a known d.c. C is adjusted until the a.c. is the same whether the coil is switched in or not. When this is done,

$$X_L = X_C/2,$$

so $\quad L_x = 1/8\pi^2 f^2 C \quad$ [C in farads]

At 50 c/s this simplifies to $\quad L_x \simeq 5/C \quad$ [C in μF]

This result is unaffected by r_x (which is not revealed by this method), but it is necessary for the voltage across C to remain constant—a condition that can be met by seeing that the impedance of the measuring circuit is small compared with that of C. For high values of L this is usually easy to arrange, even if (as in Fig. 9.19) the equality of a.c. is observed in terms of the voltage drop across a suitable resistance R, either because an a.c. milliammeter is not available, or because it would be affected by the d.c. if used. A valve or metal-rectifier voltmeter can be used, provided that any d.c. is blocked by a capacitor.

Alternatively, if the circuit impedance is made very *large* compared with that of C it keeps the *current* constant, and adjustment of C must be such as will lead to no change of *voltage* across C when the switch is operated.

In neither case is it necessary actually to measure any current or voltage in order to determine L_x, but because the inductance of an iron-cored coil depends on the current through it (a.c. and d.c.) it is desirable to adjust this to a known amount during the measurement.

This method is not recommended when the current is such as to result in much iron-core distortion.

9.21. BRIDGE METHODS FOR INDUCTANCE

For reasons already mentioned (Sec. 9.18) there is seldom much point in attempting high accuracy in measuring iron-core inductance, and one of the foregoing methods, or a bridge of moderate accuracy, is sufficient. For iron-cored coils of low resistance, the 50-c/s bridge described in Sec. 7.15 is, if rather rough, at least ready. Most of the recognized inductance bridges, such as Maxwell, Owen, and inductometer, can, if provided with sufficiently high ranges, be used for iron-cored coils, and can usually be arranged for superimposing d.c. if required. The Hay bridge is perhaps the most suitable for the particular duty of measuring iron-cored coils with some precision, and a reference to constructional details is given in Sec. 7.16. In all bridges used for iron-cored coils a tuned detector (Sec. 5.23) is almost essential. The more accurate equipment can be used not only for measuring the inductance of iron-cored coils but also for tests on core materials. This work is too specialized to describe in detail here, and reference should be made to B.S. 933. Some notes on an Owen bridge for the purpose are contained in *Muirhead Technique*, July 1949; and N. H. Crowhurst (*E.E.*, Oct. 1951) deals with the analysis of iron-core losses using a Maxwell bridge and oscilloscope. The use of a bridged-T network is discussed by J. K. Choudhury and P. C. Sen (*E. & R.E.*, Nov. 1959).

Coming now to air-cored (including iron-dust cored) coils, usually quite a standard of accuracy is wanted, because they are used for tuning or filtering specific frequencies. The Maxwell and Owen bridges are the most often chosen; the Maxwell chiefly because it can be used also for capacitance, but although simple it has to be carefully designed (Sec. 7.12). For high accuracy there is probably nothing to equal a bridge built around the standard mutual inductometer (Sec. 6.10 and 7.12).

The formulae for these various bridges are given in Fig. 7.13, *e* to *h*. Detailed instructions for the use of bought bridges are provided by the makers. The following are some general points on inductance-bridge operation.

Obviously it is necessary to avoid magnetic coupling to or from the coil under test. This is especially so when a Campbell inductometer is used, because of its large unscreened coils. Fortunately coupling falls off very rapidly with distance, and 2–3 ft of spacing is generally enough even for precise work. The leads should be rigid and about an inch apart. Residual impedance in the L_x arm, including these leads, is eliminated by taking the difference between readings with the coil in circuit and with its own terminals short-circuited. If there is any doubt about magnetic coupling to the bridge being negligible, the connections to the coil should be reversed. Any difference between the readings should be very slight, and their average taken as the answer. The subject of connection errors in inductance measurement is treated by J. F. Hersh (*G.R.E.*, Oct. 1960).

A more troublesome sort of coupling is to neighbouring metal, in which eddy currents are induced. In fact, a.f. measurements on r.f. coils in cans are misleading, unless the coils are toroidally wound, or have closed magnetic cores, because the effect of the screens on inductance as well as on resistance varies with frequency. The inductance of a coil also varies with frequency owing to eddy currents induced in its own wire (skin effect), but this need only be considered in exceptionally precise work.

The effects of self-capacitance, and of stray capacitance generally, are usually negligible when measuring r.f. coils at a.f., but may not be when measuring large inductances. If the natural frequency of L_x tuned by its own self-capacitance plus the shunt capacitance of the L_x arm of the bridge is f_r, then the apparent inductance, as measured at a frequency f, is $L_x/[1 - (f/f_r)^2]$. So if the measurement is made at one-tenth of the self-resonant frequency, the effect of capacitance is to make the measured results 1% higher than the true L_x. At one-twentieth f_r, the correction is 0·25%, which is usually negligible.

While measured inductance, corrected for self-capacitance if necessary, varies little with frequency from zero almost up to the frequency of resonance (except when eddy currents in surrounding metal are a major influence), resistance varies greatly. In general, it increases in some sort of proportion to frequency, so that Q tends to be fairly constant; but this is not to say that Q measured at a.f. can be taken as a guide to the r.f. value. At a.f., the Q of r.f. coils is likely to be very low. This has an important practical bearing on the operation of a.f. inductance bridges, for if the series resistance r_x is greater than the reactance of the coil under test, and the controls used to obtain balance are not independent as regards r_x and L_x, it is necessary to make frequent successive adjustments, and what may seem to be a fair balance may in fact be nearly correct for r_x but well off for L_x. The notorious example is the type of Maxwell bridge (Fig. 7.13*f*) in which P and/or Q resistances are varied. If R and C are varied, they balance r_x and L_x independently and the difficulty does not arise.

As Fig. 7.13 shows, what we have been calling L_x is the series value of inductance, all causes of loss being represented by resistance r_x in series. If for any reason one wants the parallel equivalents, R_x and $L_{x(p)}$, they can be calculated from the formulae in Sec. 14.12. With the Hay bridge, their equations are much simpler than the series versions (Fig. 7.13*e*). If the Q of the coil is greater than 10, the difference between the series and parallel values of inductance is less than 1%; and the parallel resistance is practically as many times greater than the reactance as the reactance is greater than the series resistance. On the other hand, if resistance is the dominating partner, Q being, say $\frac{1}{10}$ or less, its series and parallel values are practically the same, and it is the reactance (and inductance) values that differ widely. This fact is exploited in what is called Dye's shunt method for measuring inductances much higher than the ordinary range of a bridge.

9.22. HIGH INDUCTANCE BY DYE'S SHUNT METHOD

This is illustrated in connection with a Heaviside-Campbell inductometer bridge (Fig. 9.20), to which, since it is a very accurate low-

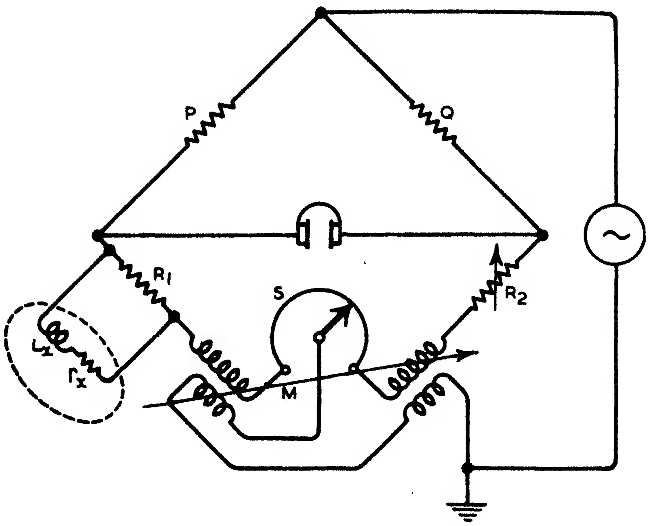


Fig. 9.20—Dye's method of using inductometer or other bridge to measure inductances much higher than its calibrated range. The same method can be used for capacitances

reading inductance bridge, it is especially applicable; but the scheme is in fact quite general, being based on the series-parallel transformation just referred to. The coil under test is shunted by a resistance R_1, much less than the reactance of the coil. The arm so formed is equivalent to nearly the same resistance in series with a reactance, and as this reactance is small compared with R_1 its inductance is within the range of a low-reading bridge. That being so, the bridge can be worked in its most accurate condition, with $P = Q$, and R_2 is made equal to R_1. Preliminary zero adjustment having been made with the coil unconnected, balance is obtained with it in position. Then

$$L_x = \frac{2\,R_1^2\,M}{r^2+(2\omega M)^2} \quad \text{and} \quad r_x = \frac{R_1^2\,r}{r^2+(2\omega M)^2} - R_1$$

where M and r are the differences in the two readings of mutual inductance and resistance respectively. If resistance balance is adjusted by the slide-wire S, *twice* the resistance traversed by the slider in making the second balance must be counted, because what is removed from one arm is added to the other. The calibration of the Cambridge instrument includes this factor of 2. If there is no slide wire, or it is inadequate, r denotes the reduction in R_2 needed to re-balance.

Since balance depends on the square of the frequency, a pure waveform must be used and the frequency must be accurately known. If the reading M turns out to be too small to be made accurately, the measurement should be repeated with a higher R_1; and vice versa.

This method, being applicable to high inductances, includes iron-cored coils, and if it is desired to measure the inductance with a known d.c. passing the arrangement can easily be made by including d.c. source and milliammeter in series. The resistance of these has to be

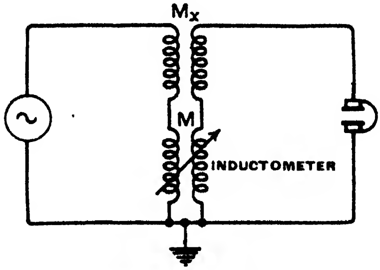


Fig. 9.21—Measurement of mutual inductance by direct balancing against mutual inductometer

deducted from the measured r_x if that is required accurately. Most of the current returns through R_1, which should be capable of standing it; but part flows through the other arms, and if it is sufficient to be objectionable in the a.c. source or detector it can be blocked from them by large series capacitances.

A wide range of capacitance can be measured by the same method, if it is connected across R_2 instead of R_1. Any resistance adjustment apart from S is of course made by R_1.

$$C_x = \frac{2M}{R_2^2}\left(1 + \frac{r^2}{(2\omega M)^2}\right)$$

Unless the capacitance is large, or has a large loss tangent, the second term in the bracket can be neglected. The method is not ideal for accurate measurement of loss tangent, but here is the formula:

$$\tan\delta = \frac{r(1 - r/R_2)}{2\omega M} - \frac{2\omega M}{R_2}$$

All the above formulae can be adapted for other types of bridge by substituting L for $2M$, L being the measured series inductance.

9.23. MUTUAL INDUCTANCE

Mutual inductance is, of course, measurable directly by an inductometer in the simple manner shown in Fig. 9.21, which needs no comment. Or it can be found by measuring the self-inductance of the primary and secondary coils connected in series, (i) when the mutual inductance between them increases the self-inductances of the whole, and (ii) with one coil reversed. Then, if these two values are L_1 and L_2 respectively,

$$M = (L_1 - L_2)/4$$

and the coefficient of coupling is

$$k = M/\sqrt{(L_P L_S)}$$

where L_P and L_S are the primary and secondary coil inductances.

9.24. IMPEDANCE

Most of the preceding sections in this chapter can be said to deal with the measurement of impedance as resistance (R) and reactance (X). These are expressed graphically as right-angle co-ordinates; e.g., in Fig. 9.18, BD represents the resistance of a coil, and DC at right angles to it represents its reactance; then BC at the angle ϕ with the horizontal or resistance axis represents its impedance, Z. Given R and X, one can easily calculate Z and ϕ (Sec. 14.12). Alternatively, however, one can measure Z and ϕ (the polar co-ordinates), from which R and X can be calculated. This is really what is done in the voltage methods described in Secs. 9.10 and 9.19, but some kinds of apparatus give Z and ϕ more directly. Constructional details of a simple example are given by N. H. Crowhurst (*E.E.*, Jan. 1949). The ranges covered are 0·2 Ω to 0·2 MΩ and $-63°$ to $+63°$, and rough indications to 0·05 Ω and 1 MΩ. The circuit is shown in Fig. 9.22. After preliminary scale adjustments, the impedance to be measured is connected and the phase control adjusted until maximum impedance is indicated by the meter. At this setting, the phase control indicates the phase angle of the

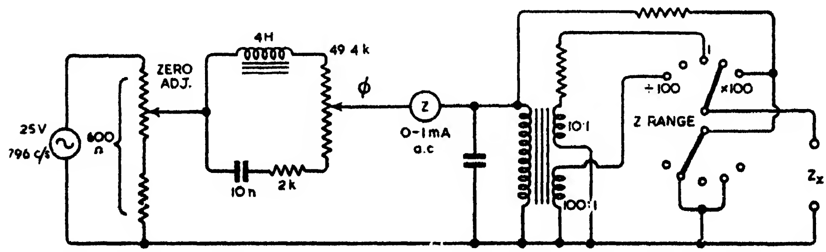


Fig. 9.22—Crowhurst's direct-reading impedance and phase-angle meter

impedance, and the meter reading (multiplied by 0·01, 1, or 100 according to the setting of the range switch) the impedance in ohms. The frequency ($5{,}000/2\pi$) is suggested by the present author to facilitate calculation. Unfortunately the method is tied to the frequency at which the phase-control circuit resonates. Another method, using even simpler apparatus, has been described by N. P. Scholes and J. E. Macfarlane (*E.T.*, March 1961).

All the methods mentioned so far reduce any system being measured to its simple series or parallel equivalent resistance and reactance—or impedance and phase angle—at the frequency of test, and make no attempt to find out the actual structure of more complicated networks. This can be done, within limits, by using the square-wave and oscilloscope bridge method referred to at the end of Sec. 7.18. To balance a bridge at all the frequencies provided by the square wave, the unknown must be matched by the standard in structure and not just in single-frequency equivalents; and the oscilloscope pictures of the unbalance provide the clues.

For the testing of transformers, loudspeaker windings, etc., see Secs. 11.3 and 11.4.

(B) MEASUREMENT AT RADIO FREQUENCIES

9.25. R.F. METHODS CLASSIFIED

There are two main classes of r.f. methods: bridge and tuned-circuit. Bridge methods, being null, are better in principle (Sec. 3.10), but although they are now being used increasingly at r.f. the difficulties are considerable (Sec. 7.17) and their technique is not included in this book. Tuned-circuit methods are comparatively simple, but nevertheless success depends very much on careful attention to detail. The best guide to this is a firm grasp of tuned-circuit theory. The necessary theory is lucidly presented, together with practical instructions, by L. Hartshorn in his *Radio-frequency Measurements* (Chapman & Hall), which is strongly recommended. It covers both tuned-circuit and bridge methods.

In all circuit and component measurements it is necessary to be clear about the relationships between resistance, capacitance, inductance, reactance, impedance, admittance, Q, phase angle, etc., but especially so at r.f. because the possibility of considering any of them separately is so much less. That being so, the following sections are grouped according to apparatus used rather than parameter measured. To avoid excessive repetition it will be assumed that if, for example, a method gives the Q and the inductance of a coil at a known frequency there is no need to explain how this information is sufficient to give also the resistance (either series or parallel), reactance, impedance, power factor, etc. This sort of thing is summarized in Secs. 14.12 to 14.17.

All the methods to be described centre on a tuned circuit. The thing to be tested is usually either the whole or part of this tuned circuit, and is measured either by substitution, using a standard of the same thing, or is deduced from other data such as voltage and frequency. The second essential element is a source of r.f. signal. The third is some sort of indicator, to enable the circuit to be tuned to resonance or to equality of frequency, and in some cases to measure the signal strength. Combinations of these essentials can be classified as three main types, according to the way in which the signal is brought into the tuned circuit (Fig. 9.23):

(*a*) In the Q-meter type, the distinguishing feature is the definiteness of the coupling from signal source to tuned circuit, whereby the amount

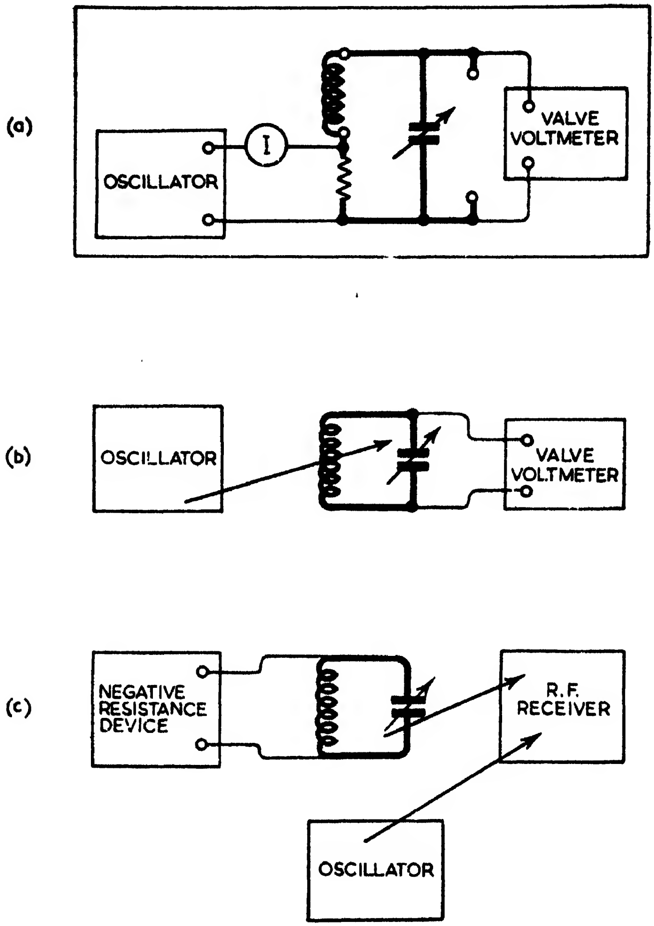


Fig. 9.23—Outline of three classes of r.f. measurement methods using a tuned circuit, shown in heavy line: (a) *direct-reading Q meter;* (b) *loose-coupled oscillator; and* (c) *negative resistance*

of signal injected is known, and one has only to measure the voltage developed across it at resonance to know its magnification and hence Q. The advantages of this are that the scale of the output voltmeter can be marked to read Q directly, and the whole apparatus can be made up as a self-contained instrument (Sec. 7.20), which is available commercially, is quick and easy to operate, and can also be used for most other tuned-circuit measurements. The disadvantages are that means must be provided to set the signal input precisely to a particular level, and that the coupling is liable to introduce appreciable error, especially at very high frequencies. The Q meter ties up a signal generator, standard variable capacitor, and valve voltmeter, all of which are very useful as independent instruments. And the ease of obtaining readings may blind one to the fact that they may differ appreciably and even substantially from the Q one wants to know.

(*b*) By sacrificing the direct-reading facility, the need for a known input signal level is avoided, and any suitable oscillator can be loosely coupled to the tuned circuit, and any valve voltmeter used as an indicator. Although there is no reason why these units should not all be permanently linked together in a self-contained system, there is also no necessity for them to be, and so they can be available for general use. This arrangement is therefore more economical but less convenient and speedy than (*a*). Obtaining results takes more time and trouble, but by taking sufficient trouble the accuracy can be very good.

(*c*) In this class of set-up, the signal source and tuned circuit are combined by making the tuned circuit the frequency-controlling part of a valve oscillator, preferably of the two-terminal type (Secs. 4.20 to 4.22). It can be set to the required frequency by comparison with a separate frequency-calibrated oscillator. An ordinary radio receiver can be used as the indicator. The advantage of this system is the extreme precision with which frequencies can be compared by the beat-note method, so it is particularly suitable for comparing very small capacitances or other measurements involving small frequency changes. It is equally precise for large frequency changes, but a little less convenient than (*a*) and (*b*) with their visual indicators because there is no indication until the frequency difference is reduced to a few kc/s.

The three foregoing types of equipment will be briefly distinguished in the following sections as: (*a*) Q meter; (*b*) loose-coupled; and (*c*) oscillating. Some kinds of measurement can be performed with two or all of these; to avoid repetition each kind will be described under one equipment and mentioned under any others that will do. It is highly instructive to make the same measurement by all three methods and compare results.

Whereas the tendency in bridge methods is to express the resistance and reactance of an impedance as series values (e.g., r_x and L_x), the parallel connection of the resonant circuits in all the foregoing systems makes it more natural to express them as parallel values. Since the parallel resistance of a tuned circuit, unlike its series equivalent, is large compared with the reactance, it is convenient to distinguish it as R_x.

9.26. MEASUREMENTS BY Q METER

Although the type diagram (Fig. 9.23*a*) shows signal injection across a very low resistance, this is not the only method; the essential thing is that a known signal voltage is introduced by some means that does not alter the tuned circuit appreciably. Before taking any Q-meter reading it is necessary to adjust the input signal to the standard level as shown

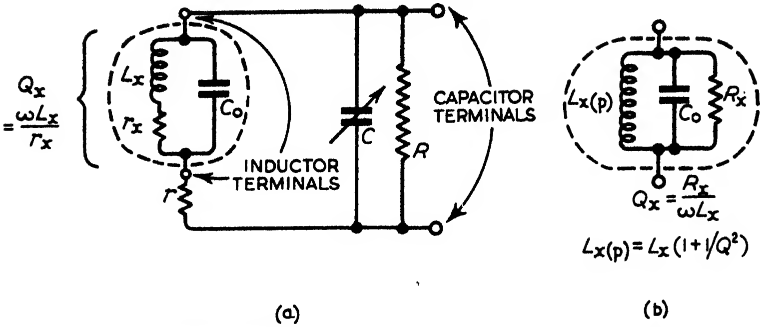


Fig. 9.24—Symbols used to denote the parts of the tuned circuit in place in a Q meter, of which r is the coupling impedance. The dotted circles enclose the alternative equivalents—series and parallel—of the inductor being measured; C is the calibrated tuning capacitor; and R the equivalent parallel resistance of the voltmeter, etc.

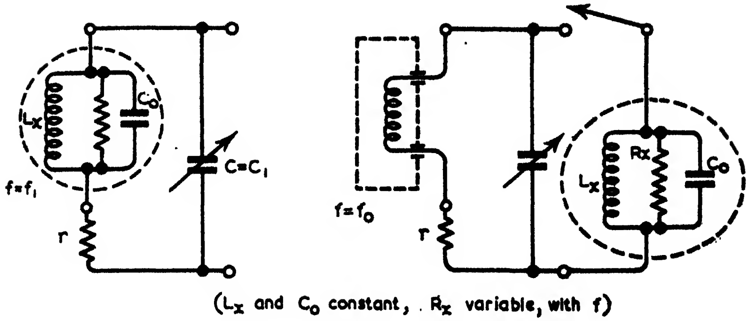
on the input meter, after the oscillator has been set to the desired frequency.

Inductors.—If the component to be measured is a coil, it is connected to the inductor terminals and tuned to resonance (indicated by a maximum on the Q scale) at the desired frequency. The scale reading is the apparent Q, which will be denoted by Q'. It differs from the true Q because of the items marked C_0, r, and R in the equivalent circuit, Fig. 9.24*a*, and also one or two other items such as the inductance of the internal part of the tuning circuit. C is the tuning capacitance, read on the instrument's capacitance scale, and should include the capacitance of the valve voltmeter, etc. R is the equivalent parallel resistance of the valve voltmeter and tuning capacitor. Usually the most important correction is for C_0, the self-capacitance of the coil, and if r, R, etc., are neglected

$$Q_x = Q'_x \left(\frac{C + C_0}{C}\right)$$

If in this and any of the following measurements r and the series equivalent of R are appreciable compared with r_x $(= \omega L_x/Q_x)$, they can be deducted from the apparent r_x and the indicated Q increased accordingly.

The next thing is to measure L_x and C_0. A good method is to set C somewhere near its maximum reading (call it C_1) and note the frequency f_1 at which the coil resonates (Fig. 9.25*a*). Then find the frequency f_0 at which the coil resonates with its own C_0 alone; that is to say, the frequency at which the coil behaves as a high resistance, being neither inductive nor capacitive. As C_1 is probably somewhere around 400 pF and C_0 of the order of 8 pF, f_0 is probably about seven times f_1. So get a coil, preferably screened, that will tune to say $7f_1$ with C near mid-scale and substitute it for the coil under test. Adjust the oscillator to resonate it. Now connect the coil under test *across* C (capacitor



(a) (b)

Fig. 9.25—Two stages in measuring self-capacitance (C_0) and inductance (L_x) by Q meter

terminals) as in Fig. 9.25*b* and note whether C has to be increased or decreased to restore resonance. If it has to be increased, increase the oscillator frequency; and vice versa. In either case try again until a frequency is found at which connecting the coil does not shift the tuning at all (Fig. 9.25*b*). This frequency is f_0.

$$\text{Then} \qquad C_0 = \frac{C_1}{(f_0/f_1)^2 - 1} \simeq \left(\frac{f_1}{f_0}\right)^2 C_1$$

$$\text{and} \qquad L_x = 1/\omega_1^2 (C_1 + C_0)$$

$$= 25{,}330/f_1^2 (C_1 + C_0) \qquad [\mu\text{H; pF; Mc/s}]$$

If L_x is small, the residual inductance of the instrument (if known) should be deducted. It is assumed that the tuning capacitor has been calibrated to include the capacitance of the valve voltmeter and the rest of the circuit; all except C_0. If this is not so, then the circuit capacitance must be measured, as described in Sec. 9.28. C_0 is assumed to be constant, but actually it varies slightly with frequency, and of course it depends on the coil's surroundings. Details of methods for correcting the indicated Q for errors due to the instrument's residuals are given by J. P. Newsome (*E.E.*, Sept. 1954; see also Feb. 1955, p. 92). The measurement of C_0 by Q meter is discussed in *E.E.*, Aug. 1956, p. 350 and Nov. 1956, p. 504.

Although connecting the self-tuned coil in parallel with the resonant circuit does not shift the tuning, its dynamic resistance R_x reduces the overall Q from Q_1 to Q_2. From these readings

$$R_x = \frac{Q_1 Q_2}{2\pi f_0 C_1 (Q_1 - Q_2)} \qquad [\text{M}\Omega\text{; pF; Mc/s}]$$

$$\text{and} \qquad Q_x = \frac{C_0 Q_1 Q_2}{C_1 (Q_1 - Q_2)}$$

These results are at the frequency of natural resonance of the coil on its own, f_0. The Q of the coil at any other frequency can be directly measured as already described, and its series resistance r_x and parallel resistance R_x at that frequency calculated as shown in Fig. 9.24*a* and *b* respectively. (Strictly, L_x is not exactly the same in these two equivalents, but unless Q is excessively small the difference is negligible.)

Tuned Circuits.—The method just described for measuring the Q of a coil at its natural frequency f_0 can be adopted for any complete tuned circuit; for example, an i.f. transformer. It is measured at its resonant frequency in exactly the same way as the self-tuned coil (Fig. 9.25*b*), and the above formulae used for finding its Q and/or dynamic resistance. The higher Q_1, the smaller the error of observation, so the coil connected to the inductor terminals should be a good one. An indispensable accessory to a Q meter (or either of its alternatives) is a set of inductors of known and stable characteristics, to cover the whole effective frequency range of the apparatus.

Resistors.—Of course it makes no difference to the Q meter whether the resistance connected to its capacitor terminals is the dynamic resistance of a resonant circuit or any other sort of resistance, as long as it is not so low as to make the second Q reading too low for accuracy. Consequently the same method and the same R_x formula can be used for measuring any resistance of about 0·02 to 2 MΩ at any r.f. If the resistor has any reactance it will shift the tuning as well as reducing the Q reading, and therefore when it has been connected C should be rotated. If resonance occurs at a new reading C_2 (the previous setting being called C_1) the parallel capacitance of the resistor is of course $C_1 - C_2$. If this turns out to be negative, then the resistor is inductive, its parallel inductance being $1/\omega^2(C_2 - C_1)$ or

$$25{,}330/f^2(C_2 - C_1) \qquad [\mu\text{H; pF; Mc/s}]$$

The reason why this simple method was not prescribed for measuring the inductance of a coil is that the value so obtained is not the true inductance but the apparent inductance, the resultant of the true inductance and any capacitance in the component.

For low resistances, see under *Capacitors*.

R.F. Chokes.—The same method is suitable for testing r.f. chokes. Ideally, these components should have an infinitely high impedance over their working range of frequency. In practice they are equivalent to a capacitance—positive at frequencies above resonance and negative below—in parallel with a high resistance. A good specimen's capacitance should not exceed about ± 4 pF nor should its resistance fall below about 0·5 MΩ. But some makes average less than 0·25 MΩ and over 8 pF, and at certain frequencies may be more like 10 kΩ and 25 pF, seriously damping and mistuning any tuned circuit in parallel with which the choke is used. The resistance and capacitance of a prospective type should therefore be measured at close intervals of frequency—close enough not to let any crevasses go unnoticed—and the results plotted. Some examples are given in *W.W.*, Nov. 1951, p. 494.

Capacitors.—The substitution method just described obviously serves for measuring any capacitance within the range of the variable in the Q meter, supplemented if necessary by known capacitances connected in parallel.

If the capacitor being measured is a very good one, that is to say its

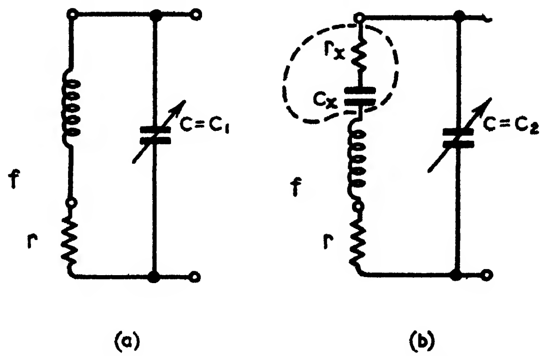


Fig. 9.26—Two stages in the series method of measuring low impedance, here typified by a large capacitance C_x with its losses represented by r_x

Q is vastly greater than that of the tuning coil and comparable perhaps with that of the tuning capacitor, its Q cannot be measured reliably; but if its loss is appreciable its equivalent parallel resistance (R_x) and its Q are indicated in the same way as for inductors:

$$R_x = \frac{Q_1Q_2}{2\pi fC_1(Q_1-Q_2)} \qquad Q_x = \frac{C_xQ_1Q_2}{C_1(Q_1-Q_2)} \qquad \tan\delta_x = \frac{1}{Q_x}$$

For capacitor-quality measurement it is particularly desirable for the Q of the tuning coil to be high. Larger capacitances can be measured, though with less accuracy, by connecting C_x in series with the tuning coil, taking care that the connections do not appreciably alter the inductance (Fig. 9.26). Then

$$C_x = \frac{C_1C_2}{(C_2-C_1)} \quad \text{and} \quad Q_x = \frac{Q_1Q_2(C_2-C_1)}{Q_1C_1-Q_2C_2}$$

Series resistance is

$$r_x = \frac{C_1Q_1-C_2Q_2}{\omega C_1C_2Q_1Q_2}$$

The accuracy of all these results tends to be poor if C_x is many times greater than C_1. The method applies to low resistances, low inductances such as the inductance of capacitors, and low impedances generally. Inductance is

$$L_x = (C_1 - C_2)/\omega^2 C_1 C_2$$

For measurement of capacitor inductance, see *Very Low Impedances*, below.

Dielectric Materials.—A sample is prepared as described in Sec. 9.13, except that it can be smaller, and even greater care is needed to ensure low resistance of the electrodes. The most suitable electrodes are tinfoil stuck to each side of the material with the least possible trace of Vaseline. Instructions are given in B.S. 903 and also in *Measurements by Q Meter* (Sec. 7.20), which includes details of a special micrometer test jig (Fig. 7.20) and edge-effect corrections. The sample is measured as a capacitor by the substitution method, and the formulae already given apply:

$$C_x = C_1 - C_2 \qquad\qquad \tan \delta_x = \frac{(Q_1 - Q_2)C_1}{Q_1 Q_2 C_x}$$

The specific permittivity or dielectric constant is

$$\epsilon_s = 11{\cdot}3\, C_x\, t/A \qquad\qquad [\text{pF; cm}]$$

where t is the thickness and A the effective area of the sample between the electrodes.

Very Low Impedances.—J. P. Newsome (*E.E.*, Nov. 1955) has shown how the range of a Q meter can be extended downward to 0·003 μH at 1 Mc/s by interposing a transformer between the usual inductance terminals and the item being measured. Constructional details of a transformer with a frequency range of about 3 : 1 are given.

One example of low inductance is that of high-C capacitors. It is necessary to make the measurement well above the self-resonant frequency of the capacitor, which normally means measuring in the v.h.f. band. In a method described by J. F. Golding (*M.I.*, Dec. 1957) the capacitor is connected by means of a simple jig straight to the terminals of a Q meter working at 50 Mc/s. The reading is compared with that of the calculated inductance of a short straight wire.

Transmission Lines.—See Sec. 11.2.

The alternative methods of measuring Q by capacitance variation and by frequency variation, described in the next section, can also be performed with a Q meter if means are provided for precise reading of small changes of capacitance or frequency. Most models include the former but not the latter, so the capacitance method would be preferred. For all measurements in which a small change in the capacitance required to tune to a given frequency has to be read, the methods given in Sec. 9.31 are preferable, because zero beat note can be located much more precisely than the exact peak of resonance.

9.27. "LOOSE-COUPLED" MEASUREMENTS

The difficulty about the type of instrument just considered is that the accuracy of Q measurement depends directly on maintaining the input signal at a known voltage and reading the output voltage. So two

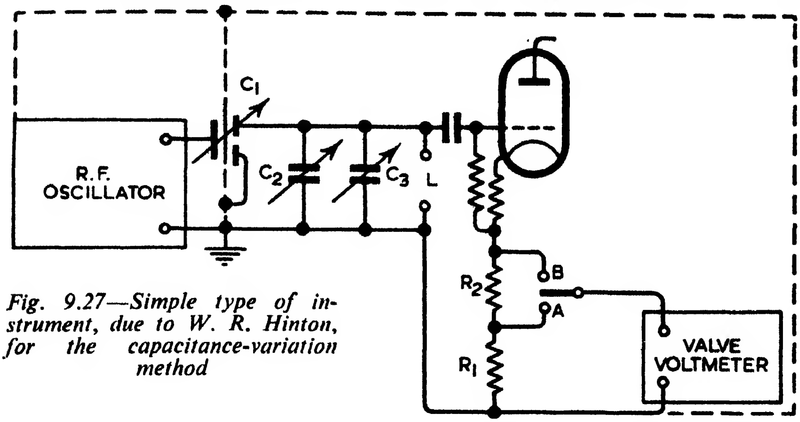


Fig. 9.27—Simple type of instrument, due to W. R. Hinton, for the capacitance-variation method

calibrations are required, and as they are quite independent (except in the " do-it-yourself " type by J. Luijckx referred to in Sec. 7.20) their errors are just as likely to add up as to counteract. So both must have good long-term stability. Calibrating and checking the valve voltmeter may not present any great difficulty, but the input signal does. And the coupling introduces an error that may be substantial enough to need correction. So unless one is fortunate enough to have a reliable professionally-made Q meter, the scheme indicated in Fig. 9.23*b* is more feasible. The input voltage must be maintained constant, but does not have to be known. Consequently, the actual voltage across the tuned circuit has no significance, and there is no need for the valve voltmeter to be calibrated in volts, but it must show voltage *ratios* accurately over at least a 1 : 0·7 range. Since there is no fixed connection from the oscillator, there is no need for the apparatus to be permanently assembled, and the oscillator, valve voltmeter, and tuning capacitor can be independent instruments. So the system is more flexible than the ready-made Q meter and can be rigged up to suit particular requirements—such as frequencies higher than those the Q meter can tackle satisfactorily. In fact, with drastic mechanical modifications it can be used right up to microwave frequencies (Sec. 12.10.) On the other hand, if it is to be used very often, and especially for measurements within a fairly limited range, there is a lot to be said for a permanent or semi-permanent set-up. Control of coupling by pushing the oscillator along the bench can be replaced by something one can do with a knob. And the residual capacitance of the tuned circuit can be kept constant. Although the calibration of the voltmeter is unimportant, it is most desirable that it should impose as little resistance on the tuned circuit as possible, as in the direct-reading Q meter, and for the same reasons. In an instrument suitable for amateur construction, due to W. R. Hinton (*E.E.*, Oct. 1951), for the capacitance-variation method of Q measurement described in the next Section, the valve voltmeter is connected via a cathode follower. Fig. 9.27 shows the circuit. C_1 is a special capacitor with a constant capacitance from the oscillator's point of view, but as a coupling capacitance it is variable from 0·02 to 0·5 pF. C_2 is the main tuning capacitor, and C_3 a small one for detuning. The coil under test is connected at "L". To avoid the necessity for the valve voltmeter to be calibrated, R_1 and R_2 are so chosen that $R_1 = 0{\cdot}707\,(R_1 + R_2)$; the only calibrations required, besides the ratio R_1/R_2, are C_2 and C_3.

Simple though such apparatus is in principle, certain requirements have to be met if the results are to be reasonably accurate. These will be better appreciated after the methods of use have been described. Direct reading of Q not being possible. one is obliged to arrive at it in some other way, usually by introducing a known variation of signal frequency, tuning capacitance, or series resistance, to reduce the voltage at resonance in a known ratio. The other readings, for example capacitance in substitution measurements, are taken in the same way as described in the previous section.

9.28. CAPACITANCE VARIATION

In all these methods the circuit is first tuned to resonance at the desired frequency, to which the oscillator has been set. The coupling to the oscillator should be adjusted to give a deflection well up the voltmeter scale. Let C_r denote the total tuning capacitance of the circuit at resonance. Now adjust the capacitance each side of C_r to divide the voltage by $\sqrt{2}$; i.e. to reduce it to 0·707 times its amount at resonance. (In Fig. 9.27, moving the switch from A to B enables this to be done at the same voltmeter reading as before.) Let ΔC denote the difference in capacitance between these two settings. Then

$$Q_x = 2C_r/\Delta C$$

The delightful simplicity of this result is due to taking $\sqrt{2}$ as the divisor. If for some reason, such as ΔC being too small to observe accurately, or the desire to take the average of several readings at different points on the resonance curve, ΔC is observed at the two points where the resonant voltage is reduced by any divisor S, the formula is

$$Q_x = \frac{2C_r\sqrt{S^2-1}}{\Delta C}$$

Note that ΔC is reckoned between the two points of reduced voltage, shown in Fig. 9.28. Although it is true that taking the capacitance difference between either of these points and the resonance peak would get rid of the factor 2 in the equation, this is not done, because the exact setting is much easier to locate on the slopes of the resonance curve than at the peak.

A vital assumption is that the frequency of the signal, and the voltage picked up by the tuned circuit (in series) or the current passed through it from the source (in parallel), are the same during both readings. To ensure that this is so, the oscillator must be stable enough for neither amplitude nor frequency to drift appreciably during the period of test. And as it must not be appreciably affected by tuning the test circuit through resonance, the coupling must be very loose. That necessitates a fairly powerful oscillator and sensitive valve voltmeter. Fortunately, circuits of lowest Q, needing closest coupling to give sufficient reading, have least effect on the oscillator and are themselves least affected. To ensure that signal amplitude is not altered by any change in the circuit or its surroundings, the coupling should be confined to a definite channel. Unauthorized coupling should be eliminated by enclosing the oscillator, the valve voltmeter and

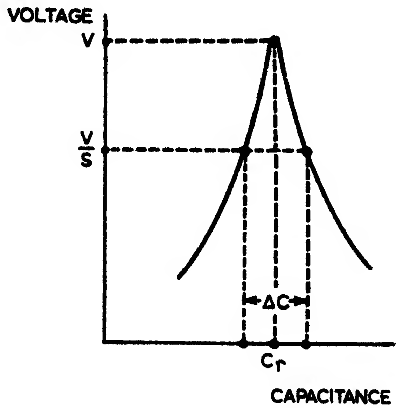


Fig. 9.28—Showing the values measured in the capacitance-variation method of measuring Q, etc.

its leads, and the tuning capacitor in earthed screens. Of course if the coupling is to be inductive the oscillator coil must not be totally enclosed, but for very precise work it should be inside a wire cage in which every wire is earthed but none makes a completely closed circuit. The coil being tested must be placed far enough from the capacitor and voltmeter screens to avoid influence therefrom, and the connections must be rigid. One should take care to occupy the same position oneself during both readings. Inductive coupling is not readily workable if either the oscillator or the coil being tested is totally enclosed, and it is not very easily controlled. The alternative is an extremely small " top-end " variable capacitance, as in Fig. 9.27.

Next, it is clear that there must be at least two capacitance controls: one to set C_r to anything up to about 500 pF, and another to observe ΔC. With high-Q v.h.f. tuning circuits, ΔC may be as low as 0·3 pF, so would have to be readable within about 0·01 pF. It is not difficult to make a low-reading capacitor, using a small unconnected moving plate overlapping two fixed ones, and well screened from the hand. A method of calibration is given in Sec. 9.34.

Details of suitable apparatus and how to set it up so as to minimize error are given more fully in Hartshorn's *R.F. Measurements*.

Assuming the apparatus is available, the only complication is in arriving at C_r correctly. It comprises the total capacitance of the main variable (which is not necessarily exactly the same as its scale reading—

see Sec. 6.6), the capacitance of the low-reading variable when set to its zero mark (preferably at mid-scale), the capacitance contributed by the voltmeter and general circuit stray, and the self-capacitance of the coil, C_0. What one wants to know is how much to add to the main capacitor reading to give C_T. Let us call it C_a (Fig. 9.29). This information is needed also for finding the inductance of the coil, L_X.

Set the main capacitor to a reading C_1, near its minimum (but preferably not too near, for the calibration may not be at its best over the top and bottom 5% of the scale) and adjust the oscillator frequency to resonate. Next, halve the frequency of the oscillator. This can be done most precisely by tuning in to its first frequency on a simple oscillating receiver (or one with a b.f. oscillator); then the second frequency is reached when its second harmonic is exactly tuned in. Note the reading C_2 where the tuned circuit resonates to the half-

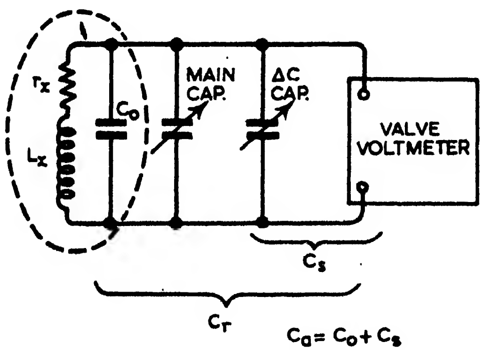


Fig. 9.29—Analysis of capacitances in capacitance-variation method of measurement

frequency (and where, incidentally, the total circuit capacitance is four times its original amount). Then

$$C_a = (C_2 - 4C_1)/3$$

The measurement can be made with any frequency ratio; the reason for choosing 2 : 1 is that it can be established exactly by the harmonic method, without having to measure any actual frequency. The general formula, if $f_2 = f_1/n$, is

$$C_a = (C_2 - n^2C_1)/(n^2 - 1)$$

If desired (and especially if the capacitor calibration is not good enough for the readings C_1 and C_2 to be highly reliable) a number of readings can be taken at any frequencies, harmonically related or otherwise, and plotted against $1/f^2$ as in Fig. 9.30. Irregular errors in the readings, shown up as departures from exact alignment, tend to average out if one draws the straight line that most nearly passes through all the plotted points. C_a is given by the negative reading of C where

$1/f^2 = 0$. If C is the capacitor reading when any frequency f is tuned in,

$$L_x = 25{,}330/f^2\,(C + C_a) \qquad [\mu\text{H; pF; Mc/s}]$$

If the graph Fig. 9.30 has been drawn, $1/f^2$ and C can best be obtained from any point near the top end of the straight line.

The use of what is basically this method, for measurements of the highest precision, is described by L. Hartshorn and J. J. Denton (*Proc.*

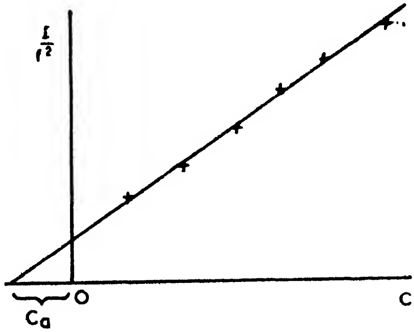


Fig. 9.30—Derivation of coil and circuit capacitance (C_a) from readings of resonant frequencies at a number of values of added capacitance C

I.E.E., Part B, July 1956) in a paper that should be studied by all concerned with accurate measurements of inductance. Crystal control is used to ensure precisely harmonic frequencies, and the capacitances are determined by switching over to a Schering bridge.

It is of course possible to derive L_x directly, without stopping at C_a on the way. The general equation is

$$L_x = \frac{\omega_2^2 - \omega_1^2}{\omega_1^2\,\omega_2^2\,(C_1 - C_2)}$$

where C_1 and C_2 are the capacitor readings corresponding to the frequencies $\omega_1/2\pi$ and $\omega_2/2\pi$ respectively. In any case, L_x includes the inductance of the leads connecting the coil to the capacitor.

If the self-capacitance C_0 of any coil is measured, it can be deducted from C_a to give the stray capacitance (C_s) of the tuning circuit with the valve voltmeter, and if this can be relied upon to be the same in subsequent use then the C_0 of any other coil measured will be known by deducting C_s from C_a measured with that coil. But it should be mentioned that the formula for C_a is a particular case of the one given in Sec. 13.3 as an example of a type to beware of. The errors in frequency can be eliminated by the harmonic method recommended, but for the capacitance readings one has to rely on the capacitor calibrations, and it must be realized that what is, say, only an error of 0·5% in a reading of several hundred pF results in a very much larger percentage error in C_a and still more in C_0. So careful measurement, using a very accurate capacitance standard, is needed to give even moderately reliable measurements of C_0*. In any case the underlying assumption that L_x and C_0 are the same at all frequencies concerned is not perfectly true, especially as regards C_0. This point is considered in Hartshorn's *R.F. Measurements*.

The foregoing procedure necessitates measuring at least one C_0 before C_s can be found; this can be done on the same lines as in Sec. 9.26, finding the frequency at which connecting the test coil in parallel with a resonant circuit makes no difference to the tuning. The methods described in that section for measuring capacitor and dielectric properties and impedances generally can also be used, subject to the different procedure for determining Q_x. There is no need, however, actually to calculate the Q before and after connecting the item to be tested across the tuning capacitor; it is sufficient to observe the values of ΔC before and after—call them ΔC_1 and ΔC_2. The formula simplifies to

$$Q_x = \frac{1}{\tan \delta_x}$$

$$= \frac{2C_x\sqrt{S^2 - 1}}{\Delta C_2 - \Delta C_1}$$

C_x is measured in the usual way, by substitution; and as usual $\sqrt{(S^2-1)}$ disappears if the $1/\sqrt{2}$ points on the resonance curve are used. Note that C_r, being the same for both readings, cancels out, removing the main objection to the capacitance-variation method. The apparatus and procedure for measuring dielectric properties in the frequency range 10 kc/s to 100 Mc/s are described by Hartshorn in his *R.F. Measurements*, and more fully with W. H. Ward in a paper† that is a model of its kind. The same formula applies to complete tuned circuits or any other impedances tested by connecting them across a resonant circuit, and in all cases the parallel resistance is given by

$$R_x = \frac{\sqrt{S^2-1}}{\pi f(\Delta C_2 - \Delta C_1)} \qquad [\text{M}\Omega;\ \text{pF};\ \text{Mc/s}]$$

The before-and-after voltage readings at resonance, V_1 and V_2, can be used as a check, for the corresponding dynamic resistances, R_1 and R_2, are proportional to them, and $V_1\Delta C_1 = V_2\Delta C_2$.

9.29. FREQUENCY VARIATION

If, instead of keeping f constant and shifting C to the slopes of the resonance curve in the capacitance-variation method of measuring Q, the tuning capacitance is kept constant and the frequency of the

* See " A Method of Measuring the Self-capacitance of Coils ", by M. G. Scroggie. *W.E.*, Sept. 1933.

† *J.I.E.E.*, Nov., 1936; also *Proc. Wireless Section I.E.E.*, March 1937.

oscillator varied to find a point each side of the peak where the voltage across the circuit is divided by S (Fig. 9.31) the equation is

$$Q_x \simeq f_r\sqrt{(S^2-1)}/\Delta f$$

and in the particular case of $S=\sqrt{2}$ this reduces to

$$Q_x \simeq f_r/\Delta f$$

The "approximately equal" sign is to show that the equation is not mathematically exact, but the discrepancy is quite negligible

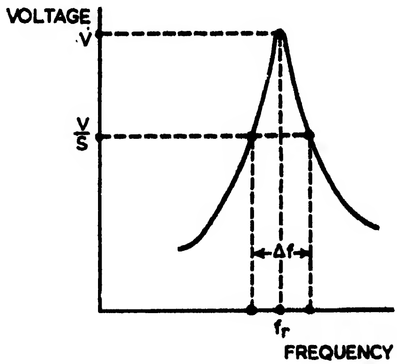


Fig. 9.31—Values measured in the frequency-variation method

unless Q_x is very low—say below 4—and in that case the assumption that it is constant over the range Δf would also break down.

The advantages of this method are that it is applicable to resonant systems with distributed capacitance, that no correction is needed for C_0, and that measuring f_r is free from the uncertainties and complications of C_T. A number of frequencies are known to 1 in 10^6 at least (Sec. 14.37)—precision enough and to spare—and one of these can either be used directly for the measurement or as a check on the oscillator calibration. For methods, see Secs. 10.9 to 10.14. As Δf is generally within the audible range, or not many times greater, the standard can be a calibrated a.f. oscillator—checked, again, by the very accurate broadcast emissions.

When the test tuned circuit is adjusted to resonance at f_r, a simple oscillating receiver is tuned to zero beat (Sec. 10.13) on the same signal. Then when frequency is shifted to the V/S points, the difference in frequency each side of f_r becomes audible as a beat note, whose frequency can be measured by comparison with the a.f. oscillator. Δf is of course the sum of the beat notes each side. E.g., if the circuit response at 1 Mc/s is reduced to 0·707 when the beat note has risen to 5,300 c/s one side and 5,100 c/s the other, Q_x is $1{,}000/(5{\cdot}3+5{\cdot}1)=96$. (All frequencies must of course be expressed in the same units, here kc/s.)

If f_r is so high in relation to Q_x that $\Delta f/2$ is above audibility, Δf can be measured in a series of steps, by adjusting the receiver periodically to shift the beat note to zero. If f_r is so high that the cumulative error due to these successive adjustments tends to be excessive, and the process tedious, a v.f. (or low r.f.) oscillator can be used to beat with the supersonic beat note and make it audible. To avoid confusing beat notes, the receiver should not be a superhet; in fact for this sort of work a single-valve receiver with self-oscillation by feedback and a pair of phones is all that is needed. For other measurements, the procedure with this apparatus is the same as described in the previous section. The last equations in that section become

$$Q_x = \frac{f_r C_x \sqrt{S^2-1}}{C_r(\Delta f_2 - \Delta f_1)} \quad \text{and } R_x = \frac{\sqrt{S^2-1}}{2\pi C_r(\Delta f_2 - \Delta f_1)} = \frac{2\pi f_1^2 L \sqrt{S^2-1}}{\Delta f_2 - \Delta f_1}$$

9.30. RESISTANCE VARIATION

This method has generally been superseded by the two foregoing, but is briefly included here because it might claim to be the classic method

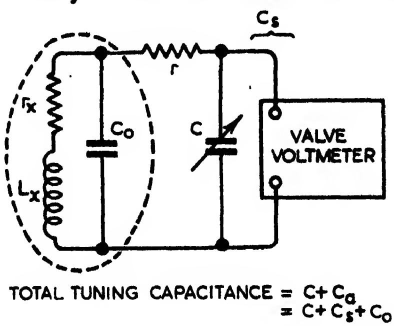


Fig. 9.32—Connections for the resistance-variation method of measuring r.f. resistance r_x; r is the resistance varied

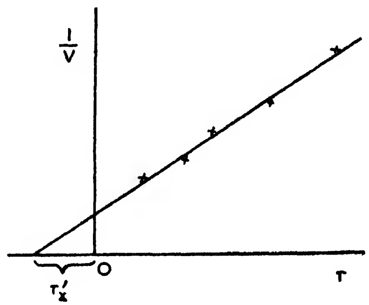


Fig. 9.33—Derivation of apparent inductor resistance (r_x') from readings of resonant voltage at a number of values of added resistance r in Fig. 9.32

for measuring r.f. resistance, and since it arrives at Q by an entirely different route it can be used as a cross-check. Fig. 9.32 shows the now familiar tuned-circuit assembly, with the coil represented inside the dotted ring by its equivalent series inductance and resistance and parallel self-capacitance, and loosely coupled to the oscillator. The only difference is the resistance r, placed in series with coil and tuning capacitor. At first it is made zero; the coil is then tuned precisely to resonance at the desired frequency and the voltmeter reading noted. With all other adjustments untouched, r is varied by inserting in turn several resistors, the r.f. resistances of which are known, and the corresponding voltages are read. It is necessary for the reactances of these resistors to be either negligible or small and equal, so they usually take the form of a short length of resistance wire, a copper wire of the same length being used as " zero " resistance. If the resistance wire is sufficiently fine the r.f. resistance can be taken as practically equal to the z.f. resistance (Table 14.8). By plotting reciprocals of voltage against resistance (Fig. 9.33) the apparent resistance of the coil, r_x', at the frequency of the test, f_r, is shown in the same manner as C_a in Fig. 9.30. Drawing the best straight line helps in the same way to average out the errors of the readings, and there is all the more need in this case, for the data are usually less accurate. Slight differences in the choice of straight line that best represents the plots make relatively large differences in r_x', and one cannot expect the result to be highly accurate. r_x' is called the *apparent resistance* of the coil, because in the test r does not carry all the current that passes through r_x. Consequently this method shares with the Q-meter and capacitance-

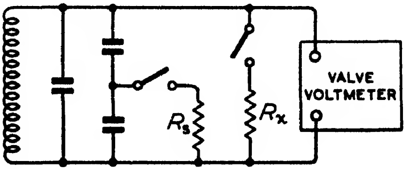


Fig. 9.34—Circuit for simple comparison method of measuring r.f. resistance

variation (but not the frequency-variation) methods the need for a C_0 correction to give the true result:

$$r_x = r_x' \left(\frac{C+C_s}{C+C_s+C_0}\right)^2 = r_x' \left(\frac{C+C_s}{C+C_a}\right)^2$$

C_a and C_s can be measured, and L_x calculated, as described in Sec. 9.28, and Q_x follows, being $2\pi f_r L_x/r_x$. So altogether one must have a good deal of time and patience to employ this method in r.f. measurements.

A method of measuring the r.f. resistance of resistors, described by G. W. A. Dummer (*W.W.*, June 1956), could also be used for the dynamic resistance of resonant circuits. R_S in Fig. 9.34 is a low-value standard resistor. When the same resonance voltage reading is obtained whichever of R_S or R_X is connected

$$R_X = R_S \left(1 + \frac{C_2}{C_1}\right)^2$$

9.31. OSCILLATOR OR DYNATRON MEASUREMENTS

In the third type of r.f. test equipment (Fig. 9.23*c*) there is no valve voltmeter; the only indicator is a receiver for (1) detecting the existence of oscillation set up in the tuned test circuit by a valve arranged to act as a controllable negative resistance; and (2) measuring its frequency by comparison with a frequency-calibrated oscillator, by the beat-note method described in Sec. 10.13. This method, and especially the variety of it known as the double-beat method, enables the test circuit to be tuned to the frequency of the oscillator with a precision limited only by the constancy and adjustability of the tuning components and the constancy of the oscillator frequency. It is an extraordinarily sensitive device, quite easily capable of detecting a change of one part in a million of capacitance, yet there is nothing delicate or expensive about the equipment. As far as frequency adjustment is concerned, it greatly surpasses either of the other systems*, so is to be preferred for such uses as measuring small capacitances.

The same apparatus can be used for measuring the dynamic resistance of the tuned circuit, and hence its Q, etc. The underlying principle is that continuous oscillation occurs when the resistance of the tuned circuit is exactly balanced or cancelled out by the negative resistance provided in the device connected to it (Fig. 9.23*c*). Since this device is connected in parallel, it is convenient to reckon in terms of parallel resistance, and therefore the resistance of the tuned circuit means its parallel or " dynamic " resistance, R_x, to which it is equivalent at resonance. The negative resistance of the device is of course also parallel resistance. If the negative resistance control is—or can be—calibrated, all one has to do is to adjust it to the point at which oscillation is just maintained (detected by the receiver) and the reading, less the minus sign, is R_x.

This is where confusion is possible. If asked which would be more effective for starting oscillation, $-10\ \text{k}\Omega$ or $-100\ \text{k}\Omega$, one might be inclined to say " $-100\ \text{k}\Omega$ ", and perhaps add " of course! " But bearing in mind that a parallel resistance of $10\ \text{k}\Omega$ represents much heavier losses than $100\ \text{k}\Omega$, one can see that $-10\ \text{k}\Omega$ must be a correspondingly more effective negative resistance to neutralize it. This follows, too, from the rule for adding resistances in parallel (Sec. 14.12), by which a combination of $+10\ \text{k}\Omega$ and $-100\ \text{k}\Omega$ is $+11{\cdot}1\ \text{k}\Omega$—still a long way off oscillation. This rather upside-down kind of addition when using parallel connections can be avoided by reckoning in conductance, as in Sec. 4.20, instead of resistance.

However, the main thing is that the apparatus under consideration can be used to measure both reactance and resistance (or susceptance and conductance) and so to specify the unknown impedance (or admittance) completely.

9.32. THE NEGATIVE RESISTOR

The only item in Fig. 9.23*c* that calls for comment is the " device " indicated so vaguely by a box. Theoretically, it is a pure variable negative resistance. Practically, it is a valve oscillator, less the frequency-determining tuned circuit. Although almost any type of oscillator circuit can be used, those that need only two connections to the tuner obviously have a great advantage over those that need tappings or coupled coils. Of these types, described in Secs. 4.20 to 4.22, one would tend to prefer whichever has least capacitance and positive conductance between its terminals (for these come across the standard variable capacitor) and the smoothest and most stable control of negative resistance over the widest range. The dynatron and transitron are very similar in these respects; the dynatron is a little

* Actually there is a method, described in Sec. 10.15, by which the frequency of a non-oscillating tuned circuit can be determined more precisely than by observing the resonant peak, but it is slower and less convenient than beat-note methods.

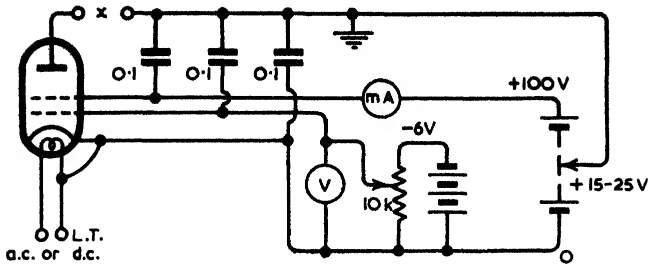


Fig. 9.35—Practical circuit dynatron as used for measuring dynamic resistance, etc.

simpler and usually throws slightly less unwanted admittance across the test circuit, but suitable types of valve are not very readily obtainable. Although exceptional valves provide resistance down to about -5 kΩ, (conductance -200 µ℧) others are limited to about -50 kΩ (-20 µ℧). Two-valve circuits (Sec. 4.22) can be devised to give a wide range with certainty, but control is generally less satisfactory. In all the types mentioned, the negative-resistance control is usually by negative bias on the control grid, and an advantage of the dynatron and transitron is that this grid forms no part of the oscillatory circuit. When the valve has such a large negative bias that current through it is almost cut off, its negative resistance is nearly infinity, and only a tuned circuit with exceptionally high R_x can be made to oscillate. At the other extreme, if R_x is very low care must be taken that reducing the bias does not make the current rise so high as to damage the valve. This is especially so with the dynatron, whose properties tend to be lost if the screen current is allowed to remain above about 7 mA for more than a second or two at a time. At very high frequencies even a high-Q circuit has relatively low R_x, so it is at these frequencies that difficulty arises.

Fig. 9.35 shows a practical dynatron circuit for measurement purposes. No particular valve is specified, as it depends on what is available (Sec. 4.20), and the tetrode symbol is intended to include pentodes with g_2 and g_3 " commoned ". One of the most important requirements is smooth control of grid bias and provision for indicating small changes in its voltage. A reliable voltmeter with a long scale is desirable, and the source of voltage ought to be very steady. Battery h.t. is shown, but a power unit can be used, provided that its resistance in the anode circuit is very low—not more than a few hundred ohms. The cathode-follower system (Sec. 4.8) is very suitable. Low resistance is needed because the negative resistance of a dynatron, unlike that of the transitron and other valve systems, is effective at z.f., and if the z.f. resistance of the source were equal to the negative resistance of the dynatron the working-point would fly to one end of the negative slope and stay there. Whatever type of negative resistance is used, obviously the source must be by-passed for the working frequency by a low-impedance capacitor connected from the low-potential x terminal by the shortest possible path to cathode. The other electrodes should have similar by-passes. It is convenient, though not compulsory, to earth this x terminal, but take care not to earth the cathode too! The other x terminal should be as close as possible to the valve. Anode voltage should be adjusted to give greatest negative conductance, consistent with smooth control—the working-point on the valve curve should be such that when oscillation starts the slope should tend to decrease; if not, oscillation, once started, jumps to a larger amplitude and the exact threshold of oscillation on the grid-bias control is not clear. At a suitable anode voltage, there is no sudden change in anode current to mark the start of oscillation; hence the need for a receiver. The fact that anode current is zero or even negative is no evidence of anything wrong. There is really no need for an anode milliammeter, but one is shown in the g_2 circuit, to warn against excessive current there. If the frequencies are suitable, broadcasting stations are economical substitutes for the beat-note oscillator, and their frequencies are known and very constant. It is usually best to use only a short length of wire as an aerial, or even stray pick-up. To check whether the dynatron oscillation is the fundamental or a harmonic, move the grid control round from the threshold point in the direction of increasing negative conductance; if the beat note continually becomes much louder it is a harmonic. The fundamental increases little for quite large movement of the grid control.

With a few obvious modifications, the foregoing remarks apply also to the transitron (*E. & R.E.*, Oct. 1958, p. 292). Suitable circuit values for conductance control are shown in Fig. 4.23. In the following sections, " dynatron " includes transitron or any other negative resistance device.

9.33. MEASUREMENT OF DYNAMIC RESISTANCE

As with the other systems dealt with in Secs. 9.26 and 9.27, the item under test is connected to the x terminals as part or whole of the tuned circuit. The tuning capacitor and the grid-bias (negative conductance) controls are adjusted so that a beat note is heard between the dynatron oscillation and the broadcast carrier or laboratory oscillator. Final adjustment of frequency should always be made when the dynatron is only *just* oscillating. The precise threshold of oscillation, which has to be observed for resistance measurement, cannot be found unless the frequency of the dynatron is made sufficiently different from that of the standard oscillator to give an audible beat note; but for resistance measurement the exact frequency is usually not very important and the frequency of the test circuit can be taken as that of the oscillator. If for any reason this is not good enough, the exact frequency can of course be measured separately by tuning the oscillator to zero beat frequency. Given a smooth bias control, it is of course very easy to make the negative resistance of the dynatron equal to the parallel resistance of the tuned circuit, R_x. The problem is to measure the negative resistance at the setting found.

One method, applicable only to the dynatron proper and the transitron with z.f. coupling, is to plot the anode-voltage/anode-current curve, the slope of which gives conductance in μA/V (micromhos) or resistance in V/mA (kilohms); but this is both tedious and of doubtful accuracy.

Another method is to calibrate the bias control by clipping known non-reactive resistors across the test tuned circuit and plotting a curve of resistance against bias-control setting. The r.f. values of such resistors are lower than the z.f., but for rough tests the z.f. values can

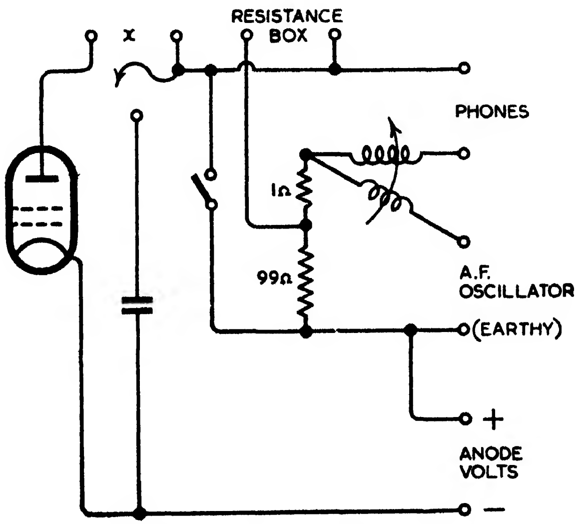


Fig. 9.36—Bridge circuit for measuring the negative resistance of a dynatron at any adjustment

be assumed up to, say, 1 Mc/s for 100 kΩ and 10 Mc/s for 10 kΩ. Of course the calibration holds good only for that particular tuned circuit at that particular frequency, and is upset by any change in any valve voltages, so it ought to be done afresh each time. However, it it is quite a useful substitution method for roughly measuring the resistance of tuned circuits, r.f. chokes, capacitors, etc. It is not directly suitable for measuring the Q of coils as such, because any coil being tested must be additional to the coil with which the system was calibrated; but it can be used for measuring the R_x of a coil at its natural frequency or as part of a complete resonant circuit. If two coils have to be connected at the same time, they should be arranged so that they do not couple with one another. If neither can be screened they must be set mutually at right angles and preferably not less than about a foot apart. The connecting leads, which should not be unnecessarily long, ought to be in position all the time, so that their capacitance is included both before and after.

The most generally useful method is by negative-resistance bridge. A simple type the author has found satisfactory is shown in Fig. 9.36 connected to a dynatron. The same method is applicable to the transitron. The two resistors form the ratio arms, so ought to be accurate and non-reactive; they can be made up as in Sec. 6.4. The

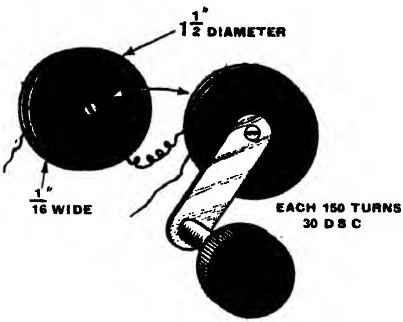


Fig. 9.37—Suitable dimensions for the phasing coils shown in Fig. 9.36

resistance box is conveniently of the decade type. A suitable frequency for the a.f. oscillator is 1 kc/s. The coupled coils are for balancing out valve capacitance, etc., and Fig. 9.37 gives suitable dimensions; the connection giving mutual inductance of the right sign is found by experiment. The switch is for shorting out the whole bridge when it is not needed, and the flying clip lead is for shorting it through the by-pass capacitor (to keep the z.f. resistance—and hence the anode voltage—constant) and alternatively shorting out the oscillatory circuit when using the bridge.

When measuring a coil, for example, it is connected to the x terminals and tuned by the standard air capacitor, assumed loss-free. The bridge switch is open and the flying lead clipped to the by-pass while the coil is being tuned and the threshold of oscillation found. When the bias control has been set right on the division between oscillation and no-oscillation, the clip is moved over to the anode terminal, shorting out the tuned circuit, and the resistance box adjusted to give exact balance. By adjusting the coupling of the bridge coils at the same time, a very sharp balance can be obtained.

For two reasons it is advisable to use an amplifier for the phones. The first is that the whole of the oscillator signal comes across the valve, and to measure the slope of (as nearly as possible) a tangent to the valve characteristic curve, the signal voltage should be kept as small as possible—at most about 1 V. As the resistance of the bridge to the oscillator is not more than 100 Ω, a step-down transformer from the oscillator is indicated. The second reason for a sensitive detector is that the resistance of the ratio arms of the bridge is necessarily low and the ratio itself is high, in order to avoid complications owing to an unduly high resistance in the anode circuit of the valve. In measuring " good " coils and circuits, the negative resistance of the valve may be several hundred kilohms, and amplification before the phones is almost essential. A step-up transformer such as a fairly high-impedance microphone-to-valve transformer, ratio between 1 : 40 and 1 : 100, may be used to advantage, and arranged so that the stray capacitance to earth is as little as possible.

Although the resistance of the bridge to anode current is small, for the utmost precision the process of setting the grid bias to the exact threshold of oscillation, and then balancing the bridge, should be repeated one or more times, in case adjustment of the resistance box has slightly shifted the critical bias adjustment. All this may sound a long business, but actually can be done very easily and quickly. Then, if R denotes the resistance of the box at balance, the negative resistance of the valve is 100 $R+99$, and numerically is a measure of R_x.

If the experiment has been done carefully there is only one correction to make; the damping caused by the terminal and anode of the valve may not be the same at the radio frequency of oscillation as it is at the audio frequency used for measuring the negative resistance of the valve. Generally the a.f. loss is completely negligible, and the r.f. loss can be measured by means of a second dynatron, the one measured being of course " dead ", with heater current off or sufficient bias to reduce negative conductance to zero. Then, if its positive conductance so measured is denoted by G_d, and G_x' denotes the measured conductance of the coil, etc., the net G_x $(=1/R_x)=G_x'-G_d$. If worked in resistances, the usual reciprocal formula must be used: $R_x=R_dR_x'/(R_d-R_x')$. This same procedure can be used for arriving at the figure for losses that cannot be measured individually but only as differences.

There is a possible error in assuming that the negative resistance of the valve is the same at all frequencies. The assumption is probably quite justifiable at any workable frequency. Care is necessary at very high frequencies, however, to prevent irregularity due to circuit resonance or absorption (Sec. 4.20).

Q, $\tan\delta$, etc., follow from R_x if L_x is known. The methods of measuring it and C_0 are so much the same as those already described in Sec. 9.28 that there should be no need to go into detail again; but the place of the valve voltmeter as a contributor of stray capacitance is taken by the dynatron. It can be measured by another dynatron as just suggested, or as part of the total circuit capacitance as described earlier. The chief practical difference is the greater ease and precision of tuning the circuit to a set frequency. One should always remember, however, to adjust the bias so that oscillation is only just maintained.

Measurement of reactance by substitution, measurement of properties of dielectric materials, and all the other measurements, can also be made by obvious adaptations of the previous methods. If R_x and C_x

are the measured values of a capacitor, $Q_x = R_x \omega C_x$. R_x is found in the same way as above, by measuring the dynamic resistance of the whole circuit, R_1 and R_2, before and after connecting the unknown; then $R_x = R_1 R_2/(R_1 - R_2)$.

9.34. MEASUREMENT OF SMALL CAPACITANCES

One thing that does call for special attention, however, is the measurement of small capacitances, and particularly the calibration of very small variable capacitors such as those required for top-end coupling and ΔC measurement in Sec. 9.28, because this can be done so much better with the class of apparatus now being considered than with the other two.

Suppose a variable capacitor to be calibrated is of the order of 10 pF. With such a small maximum, the calibration will normally be used only for differences, so any convenient point on the scale can be marked as the starting-point or zero; usually either the middle or the minimum end. This capacitor is connected to the x terminals in parallel with a coil of known inductance L and with another variable capacitance C_1, which need not be known but which must be very finely adjustable. Unless the capacitor is to be calibrated in very small steps, it is advisable to bring the total circuit capacitance up to several hundred pF, if necessary by an additional fixed or variable capacitor. With the test capacitor set to its zero mark ($C_x = 0$) and C_1 somewhere near its maximum, adjust this tuned circuit to zero beat note at some convenient frequency f. Now reduce C_1 slightly to give an audible beat note of a frequency Δf that can be compared with some standard, such as a calibrated a.f. oscillator or a broadcast standard note (Sec. 14.37). Now increase C_x, passing through zero beat and up the other side to Δf again, shifting the frequency by a total amount $2\Delta f$. The corresponding capacitance shift is

$$\Delta C_x = \frac{f\Delta f}{\pi^2 L\,(f+\Delta f)^2(f-\Delta f)^2}$$

As Δf is usually less than 1% of f, and is therefore practically negligible compared with it, the formula can be simplified to

$$\Delta C_x \simeq \frac{101{\cdot}3\Delta f}{f^3 L} \qquad [\Delta f \text{ in kc/s};\ f \text{ in Mc/s};\ L \text{ in } \mu\text{H};\ \Delta C_x \text{ in pF}]$$

Δf, f, or L can be chosen so as to make ΔC_x a round number; e.g. if Δf were 1 kc/s, L 160 μH, and f 859 kc/s, ΔC_x would be 1 pF, which would be very convenient for directly marking a scale of C_x. When the first step has been marked, C_1 is reduced through zero beat to the same beat note and the process repeated as often as desired.

The capacitor so calibrated can then be used to measure small capacitances by substitution; the tuned circuit is first adjusted to zero beat, the unknown is then introduced, and zero beat is restored by an adjustment of the calibrated variable. The same method is employed in Hunt's " Puff Box ", which contains two transistor oscillators driven by an internal 4½-V battery, and has ranges of 0 to 55 and 0 to 550 pF. Measurement is at 1·5 Mc/s. The method is also convenient for measuring the capacitance (positive or negative) of a r.f. choke. If however no calibrated capacitor is available and only a few small capacitances are to be measured, they can be found directly in terms of beat frequency, in the same manner as for calibration.

The method is effective for considerably smaller capacitances, but the ordinary zero-beat idea may have to be modified. The reason is that the beating oscillators tend to pull into complete synchronism when there is still a frequency difference between them, so that what ought to be a silent point spreads out into a silent space, and equality of frequency is difficult to locate precisely. One variation is the double-beat method (Sec. 10.13); another is to observe a change of beat note from, say, 250 to 300 c/s, as may very precisely be done by a cathode-ray tube with one pair of plates connected to the 50-c/s mains. E.g., if f is 5 Mc/s and L is 30 μH, this 50-c/s Δf corresponds to a ΔC_x of only 0·00135 pF. Of course the oscillators concerned would have to be exceptionally stable and well screened for attempting actual measurement of such a small increment.

For measuring fairly small capacitances shunted by resistance of the same order of impedance, the Richards equipment (Sec. 9.17) is convenient.

9.35. GANG-CAPACITOR MATCHING

No unusual precautions are needed for another application of the foregoing technique, namely matching the sections of a gang capacitor. Each section in turn is used to tune a suitable coil, and even very small differences in capacitance can be detected by the change in beat note. At the first (minimum-capacitance) setting the sections are equalized by means of the trimmers. Some idea of the seriousness or otherwise of mismatching at other settings can be judged without actual measurement; a beat note of a few hundred c/s, compared with zero for the section taken as the standard of reference, would generally be acceptable; but if the note goes beyond audibility altogether the mistuning effect in a receiver can be judged. The only precaution to be observed is to see that the clip lead or other means of switching from one section of the gang to another inserts the same amount of stray each time it is moved to a given section. Actual calibration can, of course, be done at the same time if desired.

(C) VALVE AND TRANSISTOR MEASUREMENT

9.36. D.C. TESTS

As mentioned in the sections on valve-testing equipment (7.21 and 7.22) there are two main techniques—d.c. meter (or static) and a.c. (or dynamic)—and these are not exactly alternatives. Both of them tell one the usual valve parameters—μ, r_a and g_m—but although a.c. tests reveal them more quickly and directly than meter tests they fail to provide characteristic curves, and when designing equipment using valves these curves are indispensable. They enable the properties of the valve to be visualized as a whole, and provide essential d.c. data. So if it is decided to use only one sort of equipment the choice must fall on d.c., which can give the whole story, though slowly. Valve a.c. bridges, on the other hand, must include meters to enable the parameters to be measured at known working-points. These parameters are sometimes called constants, but they are far from constant, and unless the working-points are specified the figures are almost meaningless. So if, in the interests of speed, a valve bridge is proposed it might as well be metered in such a way as to enable d.c. tests to be made too.

Obtaining the data for plotting characteristic curves is a straightforward process, given a sufficient number of power supplies and meters covering the appropriate ranges. The subject of power supplies is dealt with in Chapter 4. It is a great advantage if the d.c. supplies are stabilized, because otherwise an adjustment of current alters the associated voltage, and perhaps the current and voltage to other electrodes.

As there will probably not be enough meters to read the current and voltage of every electrode at once, it may be necessary either to switch a meter from one place to another, or to assume that the voltages applied to electrodes not immediately concerned are constant. Care must then be taken to avoid error due to putting meters in and out of circuit.

The voltage dropped in milliammeters or the current taken by voltmeters must be allowed for, and it is a good idea for the meters to be calibrated in this allowance as well as in their primary quantities. If, as is usual, a.c. is used for heaters, the voltmeter has to be carefully chosen, for V_h tolerances are small. Unfortunately rectifier meters are usually at their worst on this range, and other types take an inordinate amount of current. A valve voltmeter is not reliable if the heater transformer is worked under conditions that distort the waveform even slightly; for example, resistance in series with primary. A high standard of accuracy is required in the other meters too; to find the r_a of a pentode it may be necessary to measure a change of a few microamps in an anode current averaging several milliamps.

Any readings that are above the normal safe working conditions, such as those with positive grid bias, must be taken cautiously. If the power is kept on for more than a second or so the characteristics may be observed to drift and perhaps to alter permanently. When the anode current in a directly-heated valve is very heavy it alters the current distribution in the filament and results differ appreciably from those taken under working conditions, which do not allow the filament time to cool during the peaks of anode current.

A thing to beware of whenever testing valves, especially high-g_m types, is parasitic oscillation. The valve electrodes and the leads to them inevitably have capacitance and inductance, and there is always a risk of self-oscillation, usually at some very high frequency. Readings taken under such conditions are naturally highly misleading and contradictory. Oscillation can usually be detected by changes in anode current when a finger is moved to and from the grid. The risk can be minimized by not using long or closely spaced leads, and by inserting physically small resistors in series with them, close up to the valve; perhaps 1 kΩ for grid and 100 Ω for anode, provided they do not interfere with the accuracy of the tests. A reference is given in Sec. 7.21.

The valve parameters are given by the slopes of tangents to the curves at the points concerned. g_m is the slope of the anode-current/grid-voltage curve, and r_a the reciprocal of the slope of the anode-current/anode-voltage curve. μ is got by multiplying these, or by noting on the curve the ratio of the change in anode voltage to the grid voltage required to keep the anode current constant.

It is not always realized that these three are only a few of many that might be taken. For instance, in a pentode there are the g_1/g_2 mutual conductance, and $\mu_{g_1g_2}$—the change in V_{g2} required to neutralize unit change in $-V_{g1}$ so far as I_a is concerned. By the way, the standard letter symbols for valves are given in B.S. 1409.

Quite often, of course, there is no need to have any very special apparatus; the g_m of a valve can be measured—perhaps *in situ*—by applying a known increment of grid bias and noting the resulting anode current increment. Obviously the voltage actually at the anode—and in fact at all electrodes except the control grid—must be kept constant, which constitutes the main difficulty if there is appreciable resistance in the anode circuit (including the current source). The smaller the increments, the more closely the result approximates to the true slope at the point chosen, but of course the more difficult it is to read them accurately.

9.37. CATHODE-RAY TESTS

The difficulty about taking readings that would over-run the valve can be got over by projecting curves on the cathode-ray tube. The 50-c/s supply is very suitable for the purpose, as it is plentiful, is high enough in frequency to prevent the valve from being damaged in the duration of a half-cycle, and low enough not to complicate matters by capacitance currents. Fig. 9.38 shows a circuit arranged for taking anode-current/grid-voltage curves. The modifications needed for taking anode-current/anode-voltage curves are fairly obvious. V_x is the valve under test. A suitable 50-c/s voltage is applied to the grid, and also to the X plates of the cathode-ray tube. It is very likely that the voltage sweep required for the valve is not the same as is needed to produce a well-proportioned trace on the screen; so in practice a potential divider would be used and, preferably, adjusted to give a convenient scale of volts per inch or centimetre. It is not practicable to produce a vertical deflection by means of coils directly in the anode circuit, because if the number of turns were sufficient to give an adequate deflection the impedance of

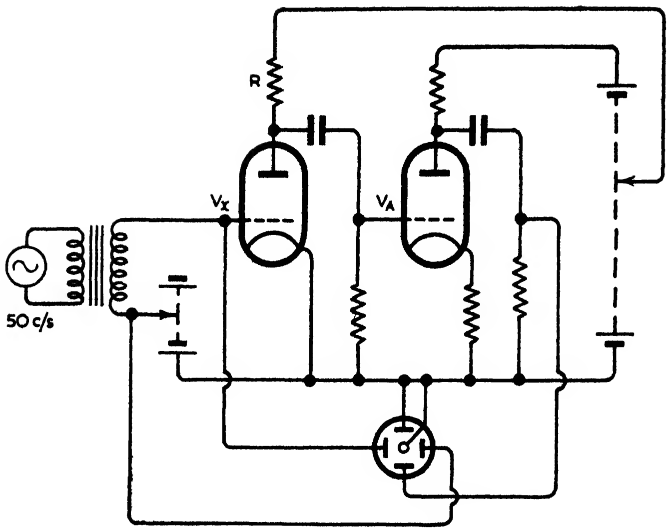


Fig. 9.38—Circuit diagram for projecting valve characteristic curves on the cathode-ray tube screen

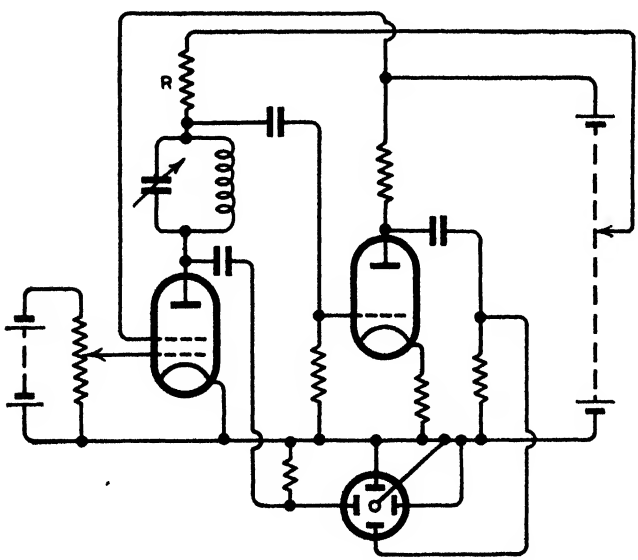


Fig. 9.39—Scheme of Fig. 9.38 modified for studying the oscillation characteristics of a dynatron

the coils would constitute a very appreciable anode circuit load. The same is true of a resistance large enough to set up a deflecting potential. So it is necessary to use amplification. The resistance *R* is as small as possible—not enough to affect the anode voltage to a serious extent—and is followed by an amplifier that is practically distortionless at 50 c/s. It should therefore be worked well within its power, and with adequate coupling capacitors—1 μF or so. It is very easy to check it by removing the input lead from R and taking it to a point on the 50-c/s potential divider that gives approximately the same signal amplitudes. If the resulting figure is a straight diagonal line, with no tendency to form a loop, there is no appreciable distortion. A similar method serves for calibration, if *R* is known.

It is possible to elaborate the scheme so as to show several curves simultaneously, by using a rapidly rotating switch, preferably

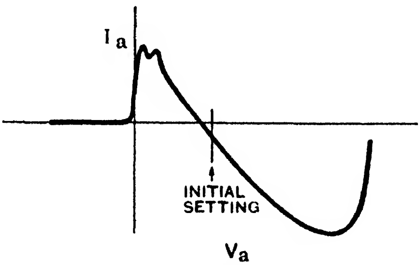


Fig. 9.40—Example of results obtained with the system of Fig. 9.39, showing over-oscillation due to insufficient negative grid bias

synchronized with the mains, to connect the anode to a number of appropriate anode voltages.

Although all this is quite interesting, the most valuable purpose of the cathode-ray apparatus is to show the behaviour of the valve under dynamic conditions. If R in Fig. 9.38 is replaced by an actual amplifier coupling, then the dynamic characteristic curves can be displayed, and the effects of altering the frequency or the nature of the coupling studied. Probably there will be enough signal developed across the coupling to make an amplifier unnecessary. A still more interesting study is self-oscillation. Suppose it is desired to see what part of the characteristic curve is swept over during oscillation; the anode current deflection is obtained as in Fig. 9.38, and either anode or grid voltage is applied to the other pair of plates. The anode oscillatory circuit is of course connected between the anode and R. Fig. 9.39 shows a circuit used for finding out the extent of the anode-current/anode-voltage curve of a dynatron over which oscillation takes place as the grid bias is adjusted, and Fig. 9.40 shows an example of the results when the bias is very much less than the oscillation threshold value.

A double-beam cathode-ray tube can be used very effectively to show two dependent variables; for example, screen current as well as anode current. R. G. Christian (*W.W.*, June 1959) uses it to superimpose X and Y axes.

References to more elaborate c.r.t. valve equipment are given at the end of Sec. 7.21.

9.38. A.C. TESTS

Deriving valve parameters such as r_a by plotting the results of meter tests as characteristic curves and measuring the slopes of those curves at the appropriate points is not only intolerably slow when many such parameters have to be measured, but is often difficult to do accurately. True, there may not be much point in achieving high accuracy in measuring items having such wide tolerances as valves, but apart altogether from instrumental errors the uncertainty in deciding which straight line has best claim to be the tangent to a curve at a given point commonly precludes even moderate accuracy. And sometimes there is a need for measuring valve parameters fairly accurately as part of some larger experiment.

Some types of valve bridge are described in Sec. 7.22, and the need for suitable power sources and metering facilities has just been mentioned. The risk of parasitic oscillation has still to be kept in mind. The standard resistors used in the bridge should be designed to be well within their ratings at the maximum current they will be required to carry.

As with the d.c. increment method, the values obtained from a valve bridge vary to some extent with the amount of valve curve swept over. No valve characteristic is perfectly straight, and the average slope for a large signal is not generally the same as for a small one. The smaller the signal the nearer the answer is to the true slope, but whereas a small d.c. signal is difficult to read accurately on a meter, in an a.c. bridge it is only a matter of using sufficient amplification before the detector. Quite a lot of amplification may be needed, not only to enable the signal amplitude to be kept small, but because some rather extreme values may have to be measured, especially with pentodes —perhaps r_a several megohms and μ several thousands. The more curved the characteristic, the less the signal that should be used. Some idea of when the signal is excessive can be judged by the amount of harmonics heard at balance. It is advisable to check the screening, etc., of the bridge by attempting to measure parameters with the cathode cold, to make sure that the measured results actually do correspond closely to infinite r_a and zero g_m.

For a bridge to measure negative resistance, see Sec. 9.33, and for detector measurements Sec. 11.24.

A very simple a.c. method of measuring r_a and μ without a bridge is described by F. E. Planer (*W.W.*, June 1945). It is analogous to the shunted-voltmeter method of measuring resistance (Sec. 9.1). Fig. 9.41*a* shows the equivalent circuit of a valve (Sec. 14.21) with an alternating voltage V_g applied to the control grid. The resulting output voltage measured across a load resistance R_L in the anode circuit is denoted by V_1. If now R_L is shunted by a capacitance C (as at *b*) the output voltage is reduced to V_1/S. From these readings r_a can be calculated:

$$\frac{1}{r_a} = \frac{2\pi f C}{\sqrt{(S^2-1)}} - \frac{1}{R_L}$$

If, as in Secs. 9.28 and 9.29, f or C is adjusted so that $S = \sqrt{2}$, this simplifies to

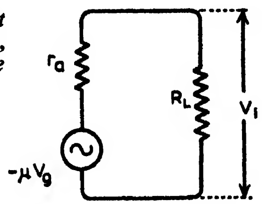
$$r_a = 1/(2\pi f C - 1/R_L)$$

The equations would be simpler still if expressed in conductances and susceptances. In any case

$$\mu = \frac{V_1}{V_g}\left(\frac{r_a + R_L}{R_L}\right)$$

C should be large enough—say not less than 3 nF—to swamp any

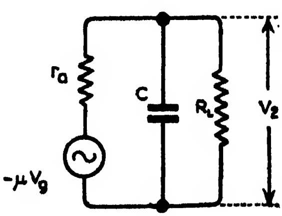
Fig. 9.41—Equivalent circuit of valve with input voltage V_g and load consisting of (a) *resistance R_L, and* (b) *R_L shunted by capacitance C. From the output readings V_1 and V_2, r_a can be found*



(a)

Table 9.3

r_a kΩ	f c/s
∞	159
1,000	161
500	162
200	167
100	175
50	191
20	239
10	318
5	478
3	690
2	955



(b)

stray capacitance across R_L. The method is at its best when r_a is not many times greater than R_L, and is not very satisfactory for r.f. pentodes. Table 9.3 gives the frequencies at which $S = \sqrt{2}$ with the values of r_a tabulated, and $R_L = 10$ kΩ and $C = 0{\cdot}1$ μF. It should be noted that if negative feedback is in use the r_a indicated is the value as modified thereby.

9.39. INPUT AND OUTPUT IMPEDANCE

Apart from the usual characteristics already discussed, one of the most important is the effect of the valve on the circuits to which it is connected. An intervalve coupling, for example, is shunted by the output impedance of the preceding valve and the input impedance of the succeeding valve. At low frequencies the former consists mainly of r_a in parallel with the capacitance of the anode to other electrodes (c_{out}), and the latter should be a practically infinite resistance in parallel with the capacitance of the grid to other electrodes (c_{in}). At high frequencies these capacitances may cause appreciable loss, which can be represented as a parallel resistance. As regards the input, the famous Miller effect (Sec. 14.19) may introduce additional capacitance and either positive or negative resistance, all varying in magnitude with frequency and the nature of the impedance in the anode circuit of the following valve. Even at a.f. this effect may be substantial. There are various other effects* which cause the input impedance to fall off steeply at v.h.f. and are the limiting factor in amplification.

The net equivalent resistance and capacitance of input, output or both, can be measured at the desired frequency by the methods described in Secs. 9.26 to 9.34,† but as the capacitances are usually small the oscillator methods (Sec. 9.31) have a special advantage, except that care is necessary to ensure that the amplitude of oscillation is kept down to what is allowable at the input of the valve being tested. Working conditions should of course be reproduced as far as possible, and it is helpful in analysing the various components of impedance to measure both with valves cold and with valves working under various conditions. Input capacitance varies somewhat, and this has an important practical bearing on constancy of tuning, particularly when a.g.c. is used.

Incidentally, all this work is clearer and simpler if the reckoning is in admittance rather than impedance.

The foregoing sections only scratch the surface of valve testing; much more information and an extensive bibliography are given in Chapter 3 of F. Langford-Smith's *Radio Designer's Handbook*, 4th edn. (Iliffe).

9.40. TRANSISTORS COMPARED WITH VALVES

Much of the four previous sections, on valve testing, applies with appropriate modifications to transistor testing. Some of the differences may now be noted.

The impedances are, in general, lower. The input impedance of a transistor, especially with common base, is particularly low—sometimes only a few ohms—in contrast with the nearly-infinite low-frequency input impedance of a valve. And instead of grid bias voltage there is base bias current. Consequently stray shunt capacitances are usually less significant than stray series inductances. As little as

* See, for example, " The Causes for the Increase of the Admittances of Modern High-frequency Amplifier Tubes on Short Waves ", by M. J. O. Strutt and A. van der Ziel. *Proc. I.R.E.*, Aug., 1938.

† For details of measurement at 1·5 to 300 Mc/s, see M. J. O. Strutt (*W.E.* Sept., 1937).

0·001 μH has been shown to cause 1 % error in test gear. And although output impedances are comparatively high, being of the same order as in pentode valves, the collector resistance to d.c. may be as little as a fraction of an ohm, and the output signal measurable in amps, so the impedances of power sources and decoupling circuits must be extremely low. The complexities of designing apparatus for accurate measurements on transistors are expounded by M. J. Gay in *Proc. I.E.E.*, Vol. 106, Part B, Supp. 15, May 1959.

Fortunately, for general laboratory purposes as distinct from transistor manufacturers' establishments or specialized research, high accuracy is seldom required. Manufacturing tolerances are so wide that it would be wasted effort. To take an example at random, the rated α_e for a certain well-established type of transistor is 40–225.

Another important difference is that input and output of a transistor have some resistance in common, so that their impedances are interdependent. In specifying transistor input impedance, for example, it is necessary to say whether the output is short- or open-circuited, and vice versa. This is especially so in the common-collector configuration.

Then transistors are much more temperature dependent than valves. I_{co} more than doubles with every 10° C rise in temperature. For this reason, and the necessarily low test circuit resistances, a transistor can very quickly be ruined. Great care must be taken never to overload it, even momentarily.

Transistors are also more frequency dependent, to such an extent with a.f. types that even 1 kc/s may be too high a test frequency at which to assume absence of high-frequency effects—at least phase shift.

Finally, provision must be made for both *pnp* and *npn* types.

9.41. TRANSISTOR MEASUREMENTS

Measurements on transistors can be divided into two classes:

(1) Sets of four parameters needed to specify them as active circuit components, for design. The so-called hybrid parameters are one such set.

(2) Separate parameters such as α, I_{co} and f_α.

Because of the very wide spread of manufacturing tolerances, the main purpose of class (1)—by a user of transistors, not a maker—is for picking out samples representing minimum, average and maximum ratings. Being small-signal parameters, they are usually measured on a special type of a.c. bridge. The purpose of class (2) may be the same, but in the smaller establishments which this book has mainly in view the object is most likely to be a check on whether a transistor is or is not in working order. For this, a fairly simple equipment is sufficient. Even more than with valves, anything like an accurate and comprehensive transistor test-gear is bound to be elaborate and correspondingly expensive, and full instructions are supplied by the makers. Only the simpler measurements will be considered here.

Probably the two most informative data for general purposes are current amplification factor (α_e) and collector leakage current (I_{coe}).

These common-emitter values are chosen not so much because that configuration is the most used, but because α_b is difficult to measure since it differs so little from 1, and I_{cob} is inconvenient to measure since it is about α_e times smaller than I_{coe}. Next, to give some idea of the possible uses for the transistor, might come $f_{\alpha e}$, the frequency at which α_e is 3 dB less than at very low (or zero) frequency. This obviously calls for a.c. operation; but if one can make do with the first two data they can be measured more simply by d.c. increment. Fig. 9.42 shows a test circuit for low-power transistors.

The power source is a two-cell lead accumulator, which is low in resistance and maintains a voltage close enough to 4 provided it is not used when just taken off a full charge or nearly due to go on again. A

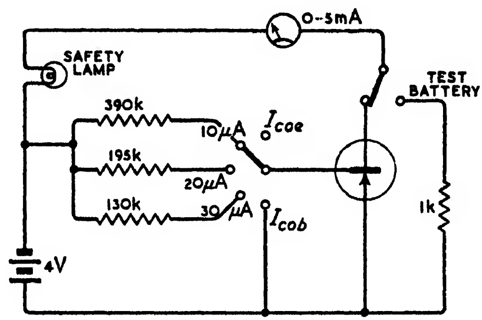


Fig. 9.42—Simple method of measuring α_e and I_{coe} of transistors

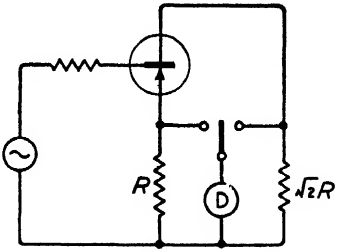
Venner two-cell silver-zinc accumulator, giving 3 V, has an even more constant discharge over its middle section; see maker's data. In calculating the resistances to pass multiples of 10 μA base current, only a few per cent error would result if the transistor input voltage were neglected, but the values shown reduce such error by allowing for typical voltages at the currents marked.

The open-circuit position of the switch gives a reading of I_{coe}, and the increments of I_c in mA as the switch is moved to the next three positions give (by multiplying them by 100) values of α_e at the corresponding mean values of I_c. Some idea of how α_e varies with I_c is thus given. The last switch position gives an approximation to I_{cob}; if it can be read on the meter there is something wrong with the transistor.

This simple arrangement can be elaborated in obvious ways, such as by using a multi-range meter and more switch positions, to test a wider range of transistor types. Slightly more elaborate instruments, but still cheaply and easily made, are those by J. N. Prewett and G. G. Yates cited in Sec. 7.23.

A.C. measurement of α and other parameters to within less than 5% calls for the carefully designed and somewhat costly equipment already referred to, based on a null method such as a bridge, or at least one in which two readings are adjusted to equality, so that the result depends on passive standards. Methods depending on the calibrations of a signal generator and a detector of some kind are generally much less accurate. However, for f_α there is only a comparison between α at some low audio frequency and at the frequency at which it is 29% less. Because of the very wide frequency range for this, it is difficult to ensure that signal generation and measurement is uniformly

Fig. 9.43—Basis of simple method of measuring $f_{\alpha b}$



effective over the whole of it. So again it is better not to depend on calibrations, even relative ones. If $f_{\alpha b}$ is measured, the low-frequency α_b can be assumed without serious error to be 1, so only the frequency at which i_c is equal to $i_e/\sqrt{2}$ need be measured. This is done in the method recommended by M. J. Gay in the paper cited, which is therefore basically as simple as Fig. 9.43. The frequency is adjusted until the detector reading is the same in both switch positions. Practical requirements add considerably to the set-up, however. And the advantage of not having to make a comparison with a lower frequency is to some extent offset by the frequency at which measurement is made ($f_{\alpha b}$) being about α_e times as high as $f_{\alpha e}$. This may be a serious consideration with v.h.f. transistors. The generator is modulated at 1 kc/s, and the detector is a diode followed by a.f. amplification and a rectifier meter, simplifying the system but limiting its sensitivity.

Given a suitable signal generator and a sensitive wide-band indicator, they can be used for measuring transistor performance at different frequencies in conventional ways, but in connecting them it should be remembered that the voltage output of the generator should be used to drive signal input current through a relatively high resistance (so as to approach the constant-current condition) and the output load should be relatively small (so as to approach the short-circuit condition).

For accurate measurement of three-terminal transistor parameters up to high frequencies—at which they are complex—there is much to be said for the Blumlein bridge (Sec. 7.10), because it responds to the direct admittance between one pair of terminals and ignores the admittances between the other two pairs. There are several commonly used sets of four parameters (Sec. 14.2); any one set completely specifies the transistor as a "black box" at the frequency used. M. J. Gay (*B.C. & E.*, June 1959) shows how the admittance parameters can be measured at frequencies from 15 kc/s to 30 Mc/s, using the Wayne-Kerr B.801 bridge (Sec. 7.13). The chief problem is to arrange the d.c. feeds and decouplings without adding to the impedances of the transistor itself. Fig. 9.44 shows Gay's circuit for measuring y_{22}—the common-base output admittance with input a.c. short-circuited. For high frequencies, the inductances of the leads must be minimized by making them of copper tape spaced 0·01 in. from a metal plate connected to the "neutral" or earthy terminal. The 0·1 mA meter, in conjunction with R_2, is for measuring the collector-to-base voltage.

J. R. James and D. J. Bradley (*E.T.*, March 1961) emphasize the usefulness of f_T plotted as a function of the d.c. working point, f_T being approximately f_1, the frequency at which the small-signal short-circuit current gain in the common-emitter configuration (i.e., α_e) is 1. They

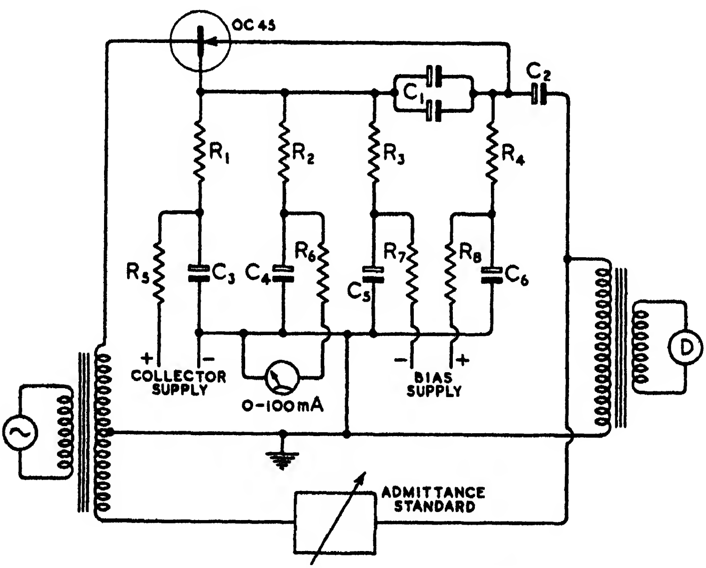


Fig. 9.44—Details of transformer bridge circuit for measuring y_{22}

show an accepted method of measuring it, employing a set-up for α_e using a r.f. signal generator at the input and a sensitive detector at the output. There is no need to go as high as f_T itself; on the basis that beyond a certain frequency α_e is inversely proportional to frequency, $f_T = f\alpha_e$, where f is the frequency at which α_e is measured.

For practical amplifier design, knowledge of power gain and noise factor are required. B. N. Harden and R. W. Smith describe the measurement of these at frequencies from 0·5 to 100 Mc/s (*E.T.*, Feb.

1961). The equipment consists of signal and noise generators and detector, connected to the transistor through special tapped input and output coils. There is a good article by C. Bayley on the theory and practice of transistor measurements (*W.W.*, July and Aug. 1961).

9.42. THERMAL PROPERTIES OF TRANSISTORS AND DIODES

One of the less obvious problems, but an important one, is to find the maximum power that a transistor or diode can safely dissipate under any particular working conditions when the maximum junction temperature is specified. A related problem is to find the thermal

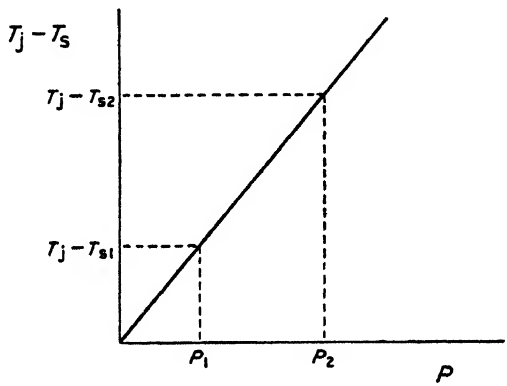


Fig. 9.45—Graph of temperature difference. between semiconductor junction and its surroundings, against power dissipated from the junction

resistance from the junction to its surroundings. This thermal resistance, denoted by θ, is the ratio of the difference in temperature between these locations to the power being dissipated:

$$\theta = \frac{T_j - T_s}{P}$$

where T_j is the temperature of the junction and T_s that of the surroundings—heat sink or ambient air, for example. θ is therefore represented by the slope of a graph of $T_j - T_s$ against P, such as Fig. 9.45. The accurate measurement of T_j when a known power is being dissipated presents difficulties, but since θ is normally linear it is possible to eliminate T_j by measuring T_s and P under two different conditions. For example, if θ is to be measured between junction and mounting stud, condition 1 could be with the transistor in free air (at temperature T_{s1}) and condition 2 with it mounted on a large metal cooling plate at temperature T_{s2} when dissipating power P_2 long enough to have reached a steady state. If the transistor is an air-cooled type, then the two different air temperatures—both, of course, below the maximum

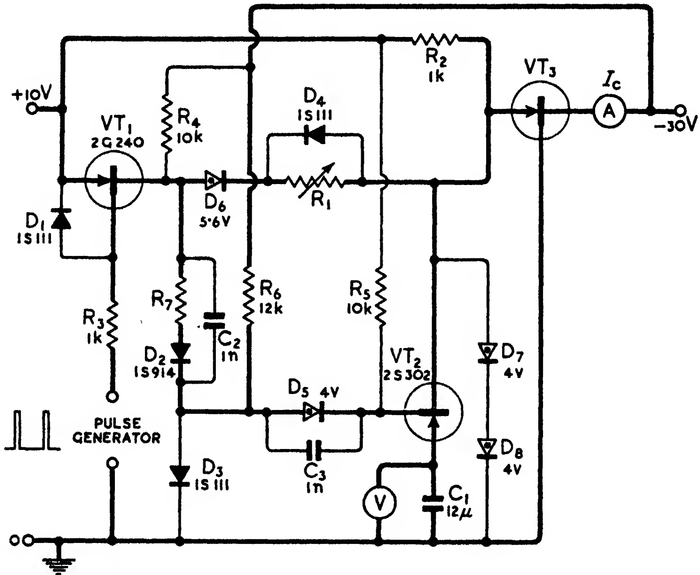


Fig. 9.46—Circuit of Texas Instruments method of measuring thermal resistance between semiconductor junction and its surroundings. The essentials are shown in heavy line; the remaining parts are refinements

allowable junction temperature!—must be got by using a controlled oven.

T_j is eliminated by taking the increment between the two conditions, as shown in Fig. 9.45:

$$\theta = \frac{T_j - T_{s1}}{P_1}$$

$$= \frac{T_j - T_{s2}}{P_2}$$

$$= \frac{T_{s1} - T_{s2}}{P_2 - P_1}$$

Although it is not necessary actually to measure T_j, it is essential that it should be the same under both conditions. This is ascertained by measuring some temperature-sensitive parameter, such as V_{be} at constant emitter current in a transistor, or forward voltage at constant current in a diode. The main difficulty is to measure it so soon after applying P_1 or P_2 that the junction has not had time to cool appreciably —say within 50 μsec. This haste tends to introduce errors through transient effects.

Fig. 9.46 shows the circuit used in a method due to Texas Instruments Ltd., which they have found to give results reproduceable to within 3%. The essential components are drawn in heavier line; the others are needed to ensure correct timing of the switching cycle and to eliminate undesired transients.

VT_3 is a transistor under test, and during 90% of the time it is passing a current adjusted by R_1 to dissipate the desired power P_1 or P_2 (equal to $I_c V_{ce}$). This condition occurs between pulses from the pulse generator, when VT_1 is saturated and VT_2 is cut off. The positive pulses, about 2,000 per second, account for the remaining 10% of each cycle; they cut off VT_1 so that only a small " metering " current defined by R_2 passes through VT_3, and at the same time open VT_2, allowing C_1 to charge to the V_{be} of VT_3, which is measured by the voltmeter V. This instrument must take negligible current, which means that it should be of the electrometer type (Sec. 5.16 - 17). The voltage so read depends on the temperature of the VT_3 base-emitter junction, which can therefore be adjusted by R_1 to be the same for both conditions.

If the actual junction temperature must be found, it can be done by open-circuiting R_1 so that only the metering current flows (and that for only 10% of the time) and the junction temperature is brought up to give the same reading on V as under load by heating the whole transistor VT_3 in a temperature-controlled oven. The power received by VT_3 due to the metering current should be compared with that needed to achieve the same T_j without the oven, to check that it is relatively negligible.

CHAPTER 10

Signal Measurements

The word " signal " in the title of this chapter is to be taken in a very broad sense, to include anything of which one can measure such quantities as voltage, power, frequency, or waveform—in contrast to the passive quantities considered in the preceding chapter.

10.1. DISTURBING EFFECTS OF METERS

In measuring current and voltage there are usually two outstandingly important considerations: the disturbing effect of the meter on whatever is being measured, and (except at z.f.) the question of waveform. It is taken for granted that the meter it is proposed to use is of a suitable type and range to indicate to the required accuracy; this has already been discussed in Chapter 5. What is now to be considered is the extent to which applying the meter actually alters the value of what is to be measured. This is mainly a matter of the impedance between the meter terminals, but sometimes there are other effects. For instance, a mains-powered valve voltmeter is likely to have a considerable capacitance between one of its terminals and earth, and if both points between which a r.f. voltage is to be measured are " live " this capacitance may upset the working considerably, even with only one terminal of the meter connected. Connecting a bulky instrument to even an a.f. amplifier may introduce feedback or hum. Avoiding such effects is mainly a matter of common sense and reasonable care, aided by experience.

Disturbance due to meter impedance most often concerns voltmeters, but sometimes has to be allowed for when measuring current, especially in low-voltage circuits such as transistor circuits. With d.c. the allowance is relatively simple, because the meter usually has a known full-scale voltage drop on all ranges. A standard value is 75 mV, but 100 mV is not uncommon in radio test-sets. The actual voltage drop in the meter at any particular reading is given by simple proportion, and if necessary can be allowed for. With a.c. it is more difficult, because the drop is in general greater and there is the awkward question of phase. Measuring heater current is a case in point; see Fig. 10.1. Even good moving-iron ammeters have a very appreciable impedance, and it would be wise to meaure the voltage on the load side of the current meter, since the extra current reading due to the voltmeter is likely to be small and easily allowed for. The accuracy required for measuring heater current or voltage is one of the highest of any measurement on valve equipment. A common tolerance on heater voltage is 7%, and on current 5% or even 3·5%, so the meter should be rather better—say accurate within 1 or 2% at worst. It should be remembered that in valves intended for parallel-connected heaters the heater current may vary appreciably among valves of the same type with the same heater voltage. And heaters for series connection should be run at the rated current, the voltages being regarded as only nominal.

At r.f. the impedance of the source may be large enough to swamp that of a current meter, but it should be checked. There are two common methods of measuring r.f. current: by thermal instrument, which demands most of one's attention to see that it is not accidentally burnt out; and by measuring the voltage drop across a known small

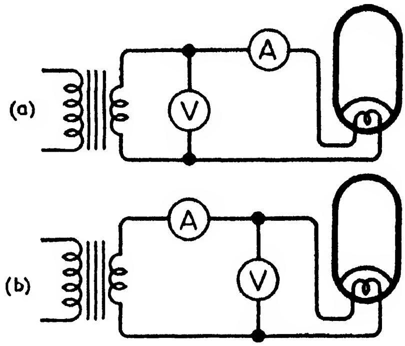


Fig. 10.1—Which method of connection is preferred depends on whether it is easier to allow for (a) *the voltage drop in the ammeter, or* (b) *the current taken by the voltmeter. In low voltage a.c. circuits like this,* b *is usually the better*

impedance by valve voltmeter, in which case the voltage taken from the circuit is known directly. The latter method is useful also for measuring very small d.c. (Sec. 9.5).

In measuring voltage, the impedance of the voltmeter is ideally infinite, and in practice ought to be many times greater than that of the source. If it is known to be n times greater, the error due to the presence of the meter can be corrected by multiplying the reading by $(n+1)/n$. In simple cases, such as in Fig. 10.2*a*, $n = R_m/R_s$; when the source is more complicated its impedance can usually be calculated by Thévenin's theorem (Sec. 14.23); for instance, in Fig. 10.2*b* the source impedance is equal to R_{ss} and R_{sp} in parallel. An alternative method, due to Bainbridge-Bell, can be used if the voltmeter has two ranges on which the unknown gives an accurate reading, and the ratio (higher to lower) of the voltmeter resistances on these two ranges (m) is known. In most instruments m is the same as the ratio of full-scale readings. Then if V_1 and V_2 are the readings on the upper and lower ranges respectively, the corrected voltage is

$$V = \frac{(m-1)V_1V_2}{mV_2 - V_1}$$

This method works best when $m = 2$. It assumes that the circuit as a whole obeys Ohm's law, so may not be satisfactory for measuring voltages in valve circuits. For the same reason it is unsuitable for use with an a.c. valve or a metal-rectifier voltmeter anywhere near the non-linear part of its characteristic, where its impedance varies rapidly with voltage. However, in the special case of a pentode resistance-coupled stage the anode voltage V_a can be measured by connecting the voltmeter first across the valve, to read V_1, and then across the load resistance, to read V_2. Provided that the h.t. voltage (V) is constant and that V_1 is above the knee of the curve, it has been shown by D. Chaplin (*W.W.*, Jan. 1954, p. 19) that

$$V_a = \frac{VV_2}{V_1 + V_2}$$

The error due to the admittance of a diode voltmeter when measuring low voltages depends on whether or not the source impedance is resonant, and it is likely to be larger than most people expect (Sec. 5.13).

With a.c. one must beware of the effects of meters not only on amplitude but also waveform. The problem arises in metering the outputs of pure waveform a.f. oscillators, for which the bridge-connected

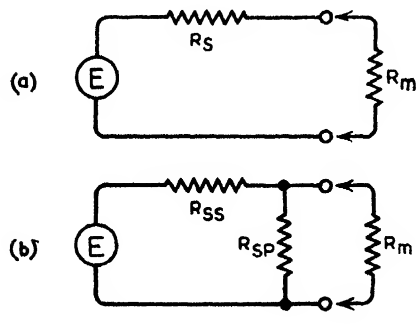


Fig. 10.2—The reduction in voltage between two points caused by connecting a voltmeter (resistance R_m) can easily be calculated if the source of voltage is equivalent (a) *to an e.m.f. in series with a known resistance R_s. In the slightly more complicated case* b, *Thevenin's theorem shows the source to be equivalent to an e.m.f. $ER_{sp}/(R_{ss} + R_{sp})$ in series with a resistance equal to R_{ss} and R_{sp} in parallel*

copper-oxide rectifier type of meter is usual. C. G. Balmain (*S.R. & R.* Feb. 1960) has given the approximate formula.

$$\frac{50\ R_o}{R_s\ (R_s + R_o)}$$

for the percentage distortion when R_o is the output resistance seen by the meter and R_s is the total series meter resistance at f.s.d.

When using an ohmmeter (or any other source of current) on transistor circuits, one must remember that transistors have a low resistance in one direction between base and either of the other electrodes. Before *in situ* measurements are made, then, the circuit must be studied to ensure that the ohmmeter is not applied with such polarity that low-resistance shunt paths exist to give misleading readings or even to damage the transistors. In the reverse polarity the current passed is usually only a few microamps and unlikely to upset the readings on the lower resistance ranges, but on the upper ranges might do so, and moreover the higher test voltages used on such ranges might well exceed the maximum reverse emitter/base rating, which in some types is as low as 7 V. This applies especially to mains-driven instruments, but some battery ohmmeters use 9 V or more. The dangers of using mains-voltage soldering irons on earthed transistor equipment have been mentioned in Sec. 2.8.

10.2. POTENTIOMETER MEASUREMENTS

A method of measuring voltage without drawing any current is by potentiometer. This is not what the radio dealer understands by a potentiometer; it is usually quite an elaborate and expensive instru-

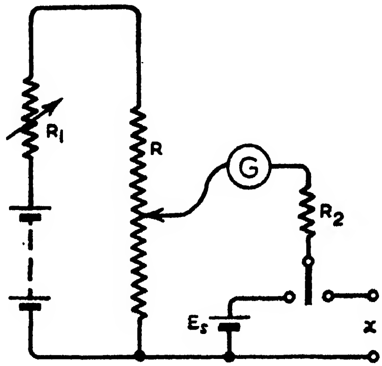


Fig. 10.3—Diagram showing principle of potentiometer, used to compare an unknown voltage at x with the standard E_s

ment, for which reason—and also the manipulation needed to obtain a reading—it is seldom used in radio laboratories. However, the principle is one that should be known, and it is quite simple.

One of the commonest uses of the potentiometer is to measure a voltage in terms of that given by a standard cell (Sec. 6.12) while it is carrying no current. In Fig. 10.3, R is the potentiometer proper; essentially it is a resistor that can be tapped at any known fraction of its whole resistance. When a steady current is passed through it, the potential between the tapping and one end is proportional to the resistance tapped off. It is not necessary to know the value of either the current or the resistance. In one method of use, a standard cell with a galvanometer in series is set to the point on the potentiometer scale marked with the standard cell voltage E_s, and the current is adjusted by R_1 until the galvanometer reads zero. The potentiometer scale then reads volts directly; an unknown voltage can be substituted for the standard cell and the tapping readjusted to bring the galvanometer reading again to zero. R_2 is to limit the current when the tapping is far off the correct adjustment, and is reduced when nearing it. Obviously it is of first importance that the current through R be constant during the measurement. The Venner silver-zinc accumulator has good characteristics for this purpose, if properly operated. In some models there is a separate tapping for the unknown, so that the standard-potential tapping can be continually monitored. And if full advantage is to be taken of the precision of a good standard cell it is clearly necessary for R to be quite an elaborate system of accurately proportioned resistors and instrument switches. For the refinements of accurate potentiometry the appropriate books (such as F. K. Harris's *Electrical Measurements*) should be consulted. Useful work can be done, however, with a simple slide wire.

The potentiometer principle can be used for comparing any voltages; e.g., for using the constant voltage across a voltage-reference tube (Sec. 6.12) to set another voltage to a desired level, without drawing current from either source.

It can also be used for a.c., but provision must be made for balancing the voltages in phase as well as amplitude, and there is no a.c. equivalent of the standard cell. Nevertheless, potentiometers are used for a wide variety of a.c. measurements; a good introduction to types and methods is given by David Owen in *A.C. Measurements* (Methuen).

10.3. OTHER NO-CURRENT VOLTAGE MEASUREMENTS

The special merit of the potentiometer is that it enables any one accurately known voltage to be used for measuring voltages over as

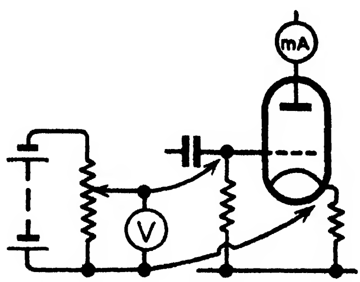


Fig. 10.4—Example of how to use a current-operated voltmeter V to measure a voltage without drawing any current

wide a range as that for which precise resistance ratios can be provided. The advantage that they are measured under no-current conditions is not exclusive, however; it can be obtained without a calibrated potentiometer. For example, suppose one has an ordinary moving-coil voltmeter with which to measure the bias voltage actually reaching the grid of a valve. It would be no good applying the voltmeter direct (Sec. 3.13). But if a source of variable voltage measured by the voltmeter is tapped on, as in Fig. 10.4, and adjusted until the anode current is the same as it was without it, V_{gk} is directly indicated. If the voltage to be measured is not already associated with a valve and milliammeter, these can be rigged up for the purpose.

The desirability of drawing no current when measuring the voltage applies to modern television e.h.t. sources, but there is some difficulty in adapting the same procedure to such high voltages. If one did happen to have a variable source of known voltage—read, say, by a current-taking voltmeter—then it would be possible to connect it in parallel with the unknown and adjust it until a galvanometer connected between it and the unknown read zero; but it would be necessary to take great care to avoid damage to the galvanometer or to oneself. If the e.h.t. were being measured under working conditions, equality of voltage could be judged without a galvanometer, by switching over from internal to measured source and adjusting the latter until the picture was the same size. But the most satisfactory solution is an electrostatic voltmeter.

10.4. SMALL VOLTAGES AND CURRENTS

Much progress has been made recently in the production of instruments for measuring small voltages and currents, especially d.c.; for information on them see Chapter 5. But choice of instrument or method should perhaps be briefly considered.

For signal voltages, meters using a thermionic diode directly applied are most useful, but in general are restricted to voltages above about 0·1 and to frequencies lower than say 300 Mc/s. In the millivolt ranges a silicon crystal rectifier working into a microammeter is usually preferred because it lacks the " splash " current which in a valve has to be balanced out and is so variable with cathode temperature. Its rectification efficiency is better at these small inputs, and its frequency range extends into Gc/s. But its long-term stability is poor, and periodic calibration checks are advisable. Alternatively, if a reliable and accurate signal generator in the same frequency band is available, the crystal detector can be used merely as a transfer instrument to compare the unknown with the known.

Within their inevitably limited frequency ranges, indicators with stabilized amplifiers are convenient, but of course the stability needed to maintain good accuracy is expensive. It may be better to sink one's capital in the aforementioned signal generator, in which case measurement down to the limits of noise can be made by using a suitable radio receiver, preferably one of the " communications " type, as the indicator for comparison. It may be necessary to alter the input connection to the receiver so that it presents a suitably high impedance to the sources of standard and unknown signals.

Sometimes it may be necessary to measure small changes in a relatively very large voltage; e.g., 0·1 V in 300 V. A simple method is to use a valve voltmeter with a very high input resistance (Sec. 5.16) with a grid blocking capacitor having negligible leakage—preferably one with polystyrene film dielectric. The time constant of the grid circuit can then be extremely long, so that quite slow changes can be measured. No intentional grid leak is used, and until the measurement is to be made the grid is shorted to earth by a switch. The technique is described by J. P. Salter (*W.W.*, Sept. 1954).

There is a commercial instrument available for this type of measurement—the Marconi Instruments TF1377—sensitive to changes as low as 1 mV in up to 500 V d.c.

In transistor circuits the need is for measuring small currents rather than voltages. This is where the clip-around milliammeter (Sec. 5.11) is so valuable, for the unsoldering of wires to insert a milliammeter is likely to do the transistor and other components no good, besides the possibility of affecting the normal functioning of the circuit. In the absence of this regrettably rather expensive instrument, one just has to do as well as one can by conventional methods, taking care to avoid overheating sensitive parts.

10.5. EFFECTS OF WAVEFORM ON METER READINGS

When measuring alternating currents and voltages one has to consider which value (peak, r.m.s., or mean) one wants to measure, and which the meter actually measures—and that is not necessarily the one in which it is calibrated. If they do happen to be the same, no difficulty arises. Nor does it even if they differ, so long as the unknown is sinusoidal, because then the three values are in known ratio (Sec. 14.11). Nearly all except special meters are calibrated in r.m.s. values. If the meter is actually a square-law type, responding naturally to r.m.s. values, this calibration holds good for all waveforms, but of course cannot be converted to peak or mean unless the factor for the particular waveform is known. The majority respond to mean or peak or some nondescript value, and the r.m.s. calibration cannot be relied upon with waveforms other than sinusoidal. If the peak or mean value of a non-sinusoidal unknown is what is actually wanted, the obvious solution is to use a type of meter responding to—and preferably calibrated in—that value (Secs. 5.2 and 5.14). But it may not be available. In such a case, and in fact in almost all work that is not confined to sine waves, it is very helpful to use an oscilloscope, which not only shows the actual waveform but enables the whole or any part of it to be measured, either by the use of calibrated shift controls if provided, or by comparison with a known voltage if they are not.

The advantage of using a meter with a true r.m.s. response is that it is the only kind that always correctly indicates the sum of two or more currents or voltages of different frequencies present together.* (If their frequency is the same the result depends on their relative phase.) "Noise"† is made up of innumerable currents, so it is normally reckoned in r.m.s. values, but there is the additional

* "Total Power." *W.W.*, March 1952.
† "Noise." *W.W.*, May and June 1952.

complication that because the currents are random the r.m.s. value (like any other) is continually fluctuating. If the noise is of the fundamental kind caused by movement of electrons, the value tends to become constant if averaged over a sufficiently long period. All meters with mechanical movements have a certain amount of inertia which slows down their response and averages the reading over an appreciable period of time; the heavier the movement the larger the time and the easier it is to read noise amplitude. The time aspect of the matter comes right to the front when the signal being measured is an irregularly fluctuating one such as speech or music. In the so-called programme meters, volume indicators and speech voltmeters, it is usual to adjust the response to the required characteristics electrically, by rectifying the signal and using circuits having suitable time constants. In this way it is possible to provide quick response to peaks, combined with a slow die-away.

10.6. MEASUREMENT OF POWER

The measurement of d.c. power, and a.c. in purely resistive loads, is covered in Secs. 5.6 and 5.7. When it is not safe to assume that the phase angle (ϕ) between voltage and current is zero, it is necessary to multiply their product by cos ϕ (the power factor); so one method is to measure the voltage and current and phase angle. This inconvenience is avoided if a suitable wattmeter is available. It seldom is, because the power to be measured in radio laboratories may be at almost any frequency, and is usually so small that the consumption of the instrument seriously complicates matters. The power consumption of equipment working off the mains, however, can be measured by means of the instrument thoughtfully supplied by the Electricity Board. The supply meter actually measures kilowatt-hours, integrating the power with time. If the apparatus under test is switched on for a definite time, everything else served by the meter being off, the average power consumed is found. It is not necessary to run the test for a long period, even though the power be small. A.c. supply meters usually contain a revolving disk visible through a window; and the number of revolutions per kWh may be specified, or if not can be derived by observing how a revolution is related to the movement of the $\frac{1}{100}$ dial, 10 revolutions of which indicate 1 kWh. To make sure that decimal points are not put in the wrong places, and to establish confidence generally, the consumption of a lamp of known wattage should be tested as a preliminary experiment.

Signal power is usually arrived at indirectly, by measuring resistance and voltage (Sec. 5.7).

10.7. PHASE DIFFERENCE

Measurement of phase difference has grown in importance with the use of amplifiers having negative feedback, and especially those for servo control. There are other applications for phase meters, such as measurement of impedances in polar co-ordinates; and although one might hardly go to the trouble or expense of acquiring one solely as an a.c. bridge detector, there is no doubt that operation of most bridges is much more pleasant if the detector is phase sensitive.

There are many methods of phase measurement, some quite simple and others extremely elaborate; a good review of them, pointing out the pros and cons of each, has been written by D. J. Collins and J. E. Smith (*E.E.*, April 1958). Comparison is usually between a reference signal, such as that applied to the input of an amplifier or bridge, and the unknown, which is the output therefrom. Some equipment enables the amplitudes to be compared too, but in others the first process is to eliminate any difference in amplitude. Most methods do not in themselves show whether the unknown leads or lags, and some device (such as provision to shift the phase slightly in a known direction) is needed. Other points to consider are the frequency range, and how much the readings are affected by waveform distortion.

The cathode-ray method, described in Sec. 9.10 in connection with capacitors, requires little besides an oscilloscope. If the angle to be measured is between two voltages, and their magnitudes are sufficient to give a good-sized ellipse on the screen, the method is quite convenient. Even so, however, it cannot be claimed as a high-precision method. Owing to the thickness of the trace, the dimensions of the ellipse cannot be measured very accurately. There is ambiguity as to which signal leads the other. And unless the waveform is pure there are liable to be errors due to harmonics, especially odd ones. The liability to error from both these causes has been examined and the results written up by F. A. Benson and A. O. Carter (*E.E.*, June 1950). Of course a double-beam oscilloscope can be used straightforwardly to indicate phase differences, by displaying both waveforms on a linear time base. A modern oscilloscope, with its fine trace and measuring facilities, is capable of fair accuracy.

To show the phase angle of a current, it is necessary to use the voltage drop across a series resistance; and usually only a small voltage can be obtained if the working conditions are not to be upset. For small voltages, amplification is needed, and it is then advisable to check—by a test on a known phase angle, such as zero with a pure resistance load—that there is no appreciable phase shift in the amplifier.

Among the simplest schemes, both in principle and practice, are those based on the fact that the sum of two alternating voltages of the same frequency depends on their phase relationship. G. de Visme (*W.W.*, Dec. 1957), after reviewing the shortcomings of other methods, described a two-valve summing circuit for use in conjunction with a valve voltmeter, and an anti-ambiguity accessory circuit. If two signals with a phase difference exceeding 180° are added together, a minimum reading can be obtained by varying the amplitude of one of them; the amplitude ratio at this minimum is equal to sin ϕ. If the phase difference is not greater than 180° it can be made so by reversing one signal. A more straightforward kind of addition, but with a less simple derivation of phase difference, is shown by H. H. Ogilvy

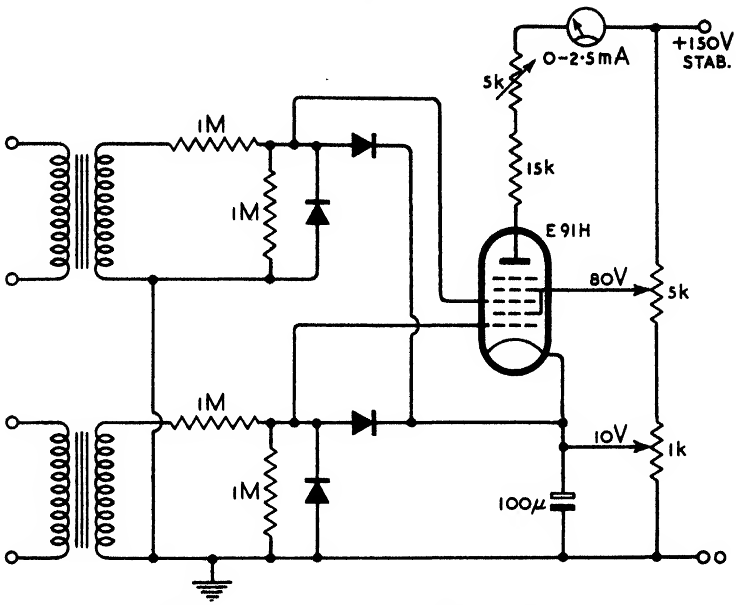


Fig. 10.5—Method of measuring the phase difference between two signals, which are both squared and applied to a frequency-changer valve

(*W.W.*, Nov. 1954), also with anti-ambiguity. Both these methods are subject to error with non-sinusoidal waveforms.

Several arrangements have been described in which the two signals are squared and applied to different grids of a mixer valve; the output is a measure of the phase difference. One such circuit was described by F. P. Moss (*E.E.*, Aug. 1954), who claimed readings independent of amplitude differences up to 20 dB and up to 20% even-harmonic distortion. Fig. 10.5 shows a simple practical circuit of German origin, using a heptode mixer. The milliammeter reading varies from zero for 5° up to 2·5 mA for 180°, and the scale is linear from 10° upwards. Another, but without squaring and therefore non-linear, is due to P. Kundu (*E.E.*, Aug. 1958).

A phase-sensitive valve voltmeter described by R. Kitai (*E.E.*, April 1957) is a complete instrument including indicator (100–0–100 μA) and variable gain, covering 20 c/s to 40 kc/s, and the phase readings are unambiguous. It is fairly elaborate, using eight double valves exclusive of stabilized power supply.

Commercially available phase meters are for the most part expensive and outside the scope of this book.

10.8. WAVEFORM EXAMINATION

One of the most valuable capabilities of an oscilloscope is for showing waveforms. The normal method of use is to apply the voltage whose waveform is to be examined to the Y plates, and a linear time base to the X plates, the time-base frequency being adjusted to a suitable fraction of the signal frequency. One-third frequency is very suitable, because it enables a complete signal cycle to be seen in any phase. The fraction must be exact if the picture is to appear stationary, and this is usually achieved by connecting the " live " Y plate to the " Sync." terminal of the time-base generator so that its sweep always begins at the same phase of the signal and each trace is an exact duplicate of the previous one. Usually one terminal of the signal source and one of the time base are " earthy ", in which case they are joined together,

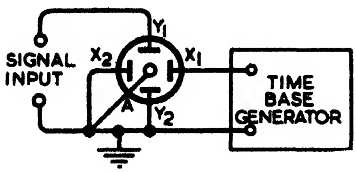


Fig. 10.6—Normal method of connecting a cathode-ray tube for examining signal waveform from a source having one terminal earthy

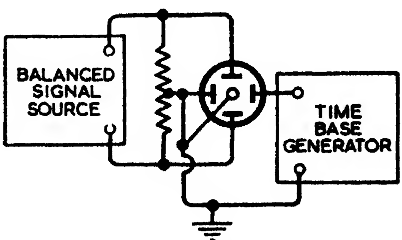


Fig. 10.7—Modification of Fig. 10.6 to suit a balanced source which has no centre or earth terminal

with the respective Y and X plates, to the c.r.t. anode, and preferably also to earth. These connections are shown diagrammatically in Fig. 10.6. Usually shift voltages are introduced between anode and the plates X_2 and Y_2. With the split beam tube, the extra plate between Y_1 and Y_2 is earthed, and both Y plates are then available for independent signals. Sometimes the signal or the time-base source is balanced to earth; i.e., both terminals are live, in opposite phase with respect to earth. If so, then of course neither plate in the pair must be connected to anode or earth; the connection must be made to the centre point of the signal source. If none is provided then one must be made, say by using a centre-tapped resistance, sufficiently high not to affect the waveform appreciably, as in Fig. 10.7. In practice, amplifiers are usually interposed between signal sources and c.r.t. plates, and in most modern instruments they have push-pull output. Their linearity is obviously of first importance in this application.

Special care must be taken with waveforms that contain very high frequencies, or these will be reduced by the shunt capacitance of the connecting leads and c.r.t. plates or of the amplifier, and the waveform thereby altered. The line-synchronizing signals in television are an example; although their fundamental frequency (in a 405-line 50-c/s system) is only 10·25 kc/s, their clear-cut shape depends on the inclusion of harmonics up to about 1 Mc/s, at which even the input capacitance of the oscilloscope alone may be an impedance of only about 10 kΩ. So it is necessary to work it from a source with a lower impedance still, and if this is not true of the actual source it is advisable to use an attenuating probe as in Fig. 5.40, followed by a wide-band amplifier to make up for the lost amplitude. This is usually a more effective use for a valve than as a cathode follower.

If an amplifier is used, the sync connection should be from its output, not only to ensure sufficient amplitude for synchronizing but also to prevent the time-base voltage from reaching and perhaps distorting the signal. In fact, even if no amplification is needed a buffer stage may have to be used for this reason; a cathode follower is very suitable.

It may be possible to do without amplification for many television waveforms, but picking out the particular section one wants is likely to be more difficult than might be expected. Suppose, for example, it is desired to examine the first odd-frame synchronizing signals. This means selecting and displaying one line out of the 405 making up one complete television waveform, and keeping it steady. Clearly all the unwanted lines must be suppressed in some way; and a phase control of exceptional precision is needed for picking out the desired line. One example of apparatus for this purpose is described by K. R. Sturley (*W.E.*, Sept. 1951), and a more elaborate one by R. Anderson and J. R. Smith in *Proc. I.E.E.*, Part IIIA, No. 19, 1952.

For analysis of waveform and measurement of distortion, see the next chapter, from Sec. 11.12.

10.9. FREQUENCY MEASUREMENT

Of all parameters, frequency is unique in the extreme accuracy of standards constantly available to all without extra charge, and the precision with which comparisons can be made. A particularly valuable feature is that one frequency can be set to a fraction or multiple of another as precisely as to equality. Information on frequency standards and associated apparatus is given in Secs. 6.13 and 14.37, and frequency meters in Sec. 7.19.

Some instruments merely have to be connected to a signal to indicate its frequency, all the work being done for one internally. There are two main kinds of these (Sec. 7.19): the frequency-discriminator type, which gives a continuous direct reading on a meter, so that varying frequency can be followed, but the accuracy is only of the few-per cent order; and the digital counter type, which shows the number of cycles occurring within an accurately measured short period of time, and can be highly accurate but is expensive. In both cases the mode of use is obvious, so we now consider only methods of comparison with some kind of frequency standard. Although not strictly within the scope of this chapter, frequency of circuit resonance will be considered, including the use of non-generating or passive standards of frequency.

The very highly accurate standards, with errors of one or less in a million, are available as radio or radio-borne signals on a few spot frequencies. Given even only one of these, it is possible to fix as

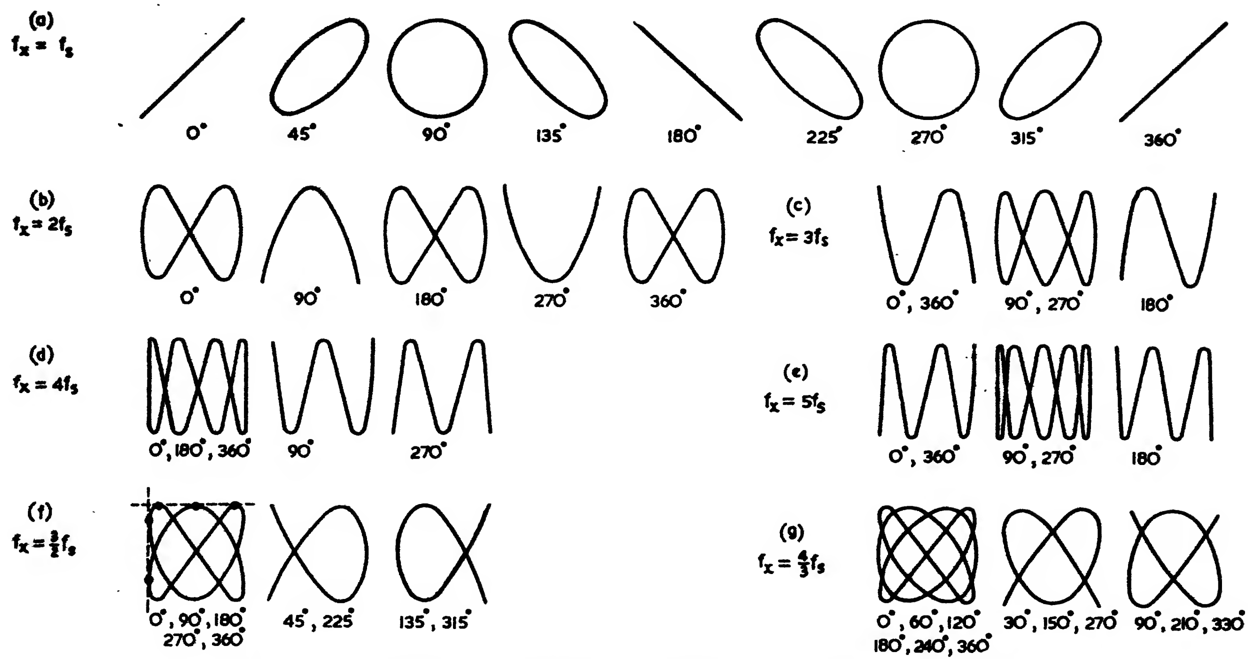


Fig. 10.8—Examples of Lissajous figures obtained on a c.r. tube when sinusoidal voltages having the frequency ratios and phase differences specified are applied to the two pairs of plates

many other frequencies as one likes, but it takes much too long to do so whenever a particular frequency is to be measured. To avoid this, there are methods employing apparatus generating fixed spot frequencies at more or less close intervals over the whole useful range and held in synchronism by the standard, so that all are known to the same degree of accuracy (Sec. 6.15). For general purposes, however, it is sufficient to calibrate the a.f. and r.f. oscillators that are needed in any laboratory, and check the calibration against the standard whenever necessary. This, then, will be the object chiefly in mind in the following sections.

10.10. FREQUENCY COMPARISON BY CATHODE-RAY TUBE

The simplest method is to connect the signal of known frequency (f_s) to the X plates and the unknown (f_x) to the Y plates. If f_x is exactly equal to f_s, and in phase with it, the trace is stationary and takes the form of a diagonal straight line. Forms corresponding to a number of other phase relationships are shown in Fig. 10.8*a*; see also Sec. 9.10. Throughout Fig. 10.8 the two signals are sinusoidal and equal in amplitude, these being the ideal conditions. With signals of other waveforms the traces are distorted; but provided the percentages of harmonics are moderate they should not be unrecognizable. Unequal amplitude makes the traces appear either flatter or narrower.

If f_x is nearly but not quite equal to f_s, there will be a constantly increasing phase difference between the two, making the trace go through all the forms shown at *a* continuously. The time taken to go through the whole lot is the time of one cycle of the frequency *difference*; e.g., if f_x is decreased and the trace moves more slowly, $f_x > f_s$.

If, now, f_x is twice f_s, patterns such as Fig. 10.8*b* are produced; if three times, *c*; and so on. They are all examples of what are called Lissajous figures, discussed in detail by Hilary Moss in a series of articles in *Electronic Engineering*, beginning June 1944; also in book form under the title *Cathode Ray Tube Traces*. The phase angles specified refer to the phase of the f_x signal at the instant when the f_s signal is at 0° (i.e., zero X deflection). The ratio f_x/f_s is found by counting the number of loops touching a horizontal tangent (shown dotted in Fig. 10.8*f*) and dividing by the number of loops touching a vertical tangent; but beware of phase angles, such as 45° and 135° in this case, that give traces with " loose ends ", for in these there are loops superimposed so that their full number is not seen.

Unless the pattern can be halted at the right phase, even a small number of loops is difficult to count, and a better method is to connect the signal of lower frequency to both pairs of plates as in Fig. 10.9. C should be chosen so that its reactance at that frequency is about R ohms; e.g., for 50 c/s, 0·1 μF and 30 kΩ would do. If the other signal is connected in series with the anode supply the result is a toothed-wheel pattern, in which the number of teeth indicates the frequency ratio, and a slight departure from exact ratio makes the wheel rotate.

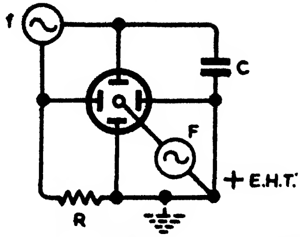


Fig. 10.9—Method of connecting two signal sources f and F so as to obtain traces such as Fig. 10.10, in which there are F/f teeth to the wheel. The reactance of C at frequency f should be approximately equal to R

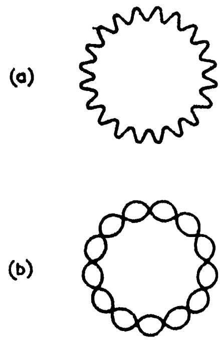


Fig. 10.10—Two traces obtained by the method shown in Fig. 10.9 Example a *shows $F = 20f$;* b, *$F = 13f/2$*

If the lower frequency is denoted by f, the higher by F, the time for one complete revolution by T, and the number of teeth by N,

$$f = \frac{F}{N} \pm \frac{1}{T}$$

E.g., in checking the mains frequency by a standard 1,000-c/s signal, a stationary 20-tooth wheel (Fig. 10.10*a*) would indicate exactly 50 c/s. But if the wheel moved one tooth's pitch in one second, T would be 20 sec, and the mains frequency $50 \pm 0{\cdot}05$ c/s. Once it has been found by experiment, as in the previous method, whether a clockwise rotation means frequency " fast " or " slow ", it is very easy to see whether the correction is + or −.

The method can be used for simple fractional ratios, as in Fig. 10.10*b*, where the ratio is 13 : 2.

If the available f signal voltage is insufficient to show teeth, it may be connected to c.r.t. grid instead of anode, and the bias adjusted to give a dotted-line effect. But beware of ambiguity; it is possible, for example, to obtain similar patterns with a frequency ratio of either 3 : 1 or 3 : 2.

10.11. USE OF AUXILIARY OSCILLATOR

With a known 50 c/s only (for example), it is possible to calibrate an oscillator at fractions of this frequency—25, 16⅔, 12½, etc.—and multiples—100, 150, 200, etc. As about 20 teeth can readily be distinguished for each centimetre of circle diameter, the whole a.f. band can be included with a tube of moderate size; but unless the frequencies are phenomenally steady at the exact multiple the only way of counting the highest numbers is to increase the frequency very gradually from some easily countable number and note every time the pattern " pulls in ". Even this requires some care, and a more reliable method is to have an auxiliary oscillator tuned to, say, 1,000 c/s by the 50 c/s; and then use this to form the circular base for the higher frequencies. Another advantage is that many new points are obtainable: from 1,000 c/s—$333\frac{1}{3}$, $166\frac{2}{3}$, $142\frac{6}{7}$, $111\frac{1}{9}$, and so on. Then the auxiliary oscillator can be shifted up another 50 c/s by the standard, and still another series obtained. There is no limit, except patience, to the number of points that can be derived from one fixed standard frequency for drawing a curve and so obtaining a continuous calibration.

The foregoing numbers are merely an example to illustrate the general principle, and should not be taken to imply that the frequency of the mains is an accurate standard. Although it has a very high

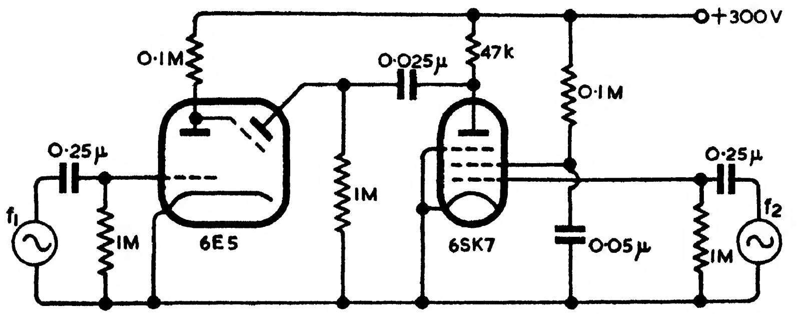


Fig. 10.11—Circuit of indicator for showing when f_2 is equal to f_1 or a multiple thereof up to about 10

long-term-average accuracy, suitable for electric clocks, appreciable fluctuations are possible. However, when the system is not overloaded it is usually well within 1% and so can be used for rough purposes.

10.12. FREQUENCY COMPARISON BY TUNING INDICATOR

Where it is not necessary to discover the frequency ratio but only to adjust frequencies to equality, there is no need for a large and expensive oscilloscope: the smaller and cheaper type of cathode-ray tube known popularly as the magic eye can serve the purpose. It is particularly useful for building into a calibrated oscillator to enable the calibration to be brought into exact agreement with some external standard by means of a pre-set control. There are many possible arrangements; the one shown in Fig. 10.11 is due to K. G. Beauchamp.* When f_1 is nearly equal to f_2 the indicator can be seen to flicker, at a frequency equal to the difference. There is a similar but less clear indication when one frequency is a whole multiple of the other, up to about 10 times.

10.13. AURAL COMPARISON

Although radio frequencies *can* be compared visually, the higher the frequency the more difficult it is to find and maintain a fixed or

* *E.E.*, May 1951. Another circuit, by H. V. Beck, appears in the Oct. 1951 issue.

slowly moving pattern, whereas aural comparison is actually easier when the frequencies to be compared are above audibility. Moreover their harmonics can be examined separately, without interference from even far stronger fundamentals. Audio frequencies too can be compared aurally, but success depends somewhat on one's sense of pitch; a musical ear is a great help. The pitch of a sound containing harmonics appears to the ear to be the frequency of the fundamental, though that be relatively weak, or even entirely absent! But some people have difficulty in telling whether a note rich in harmonics is higher or lower than another. If, however, both are reasonably pure, most individuals can adjust one frequency towards the other until first the characteristic trill and then slow beating is heard as equality is closely approached. The frequency of beating is the difference between the component frequencies. Here again, if there are strong harmonics a person with a poor sense of pitch may be unable to decide whether the beating is due to the fundamental or a harmonic.

If both signal frequencies are audible, there is no need for the signals to be combined; they can be listened to from separate loudspeakers or earpieces. If both are above audibility but their difference is within the a.f. band, the beat note can be heard, provided that the two r.f. signals are combined non-linearly. There is no need to do anything very special about this, because in practice the signals come to the phones or loudspeaker through a valve, and valves are always non-linear enough to yield some signal at the beat frequency.

A great deal of variety is possible in the arrangements for obtaining the beat note. For calibrating an oscillator against a broadcast frequency, for example, one would obviously use a suitable receiver to bring in the standard, and there would usually be no difficulty in arranging for a signal from the oscillator also to be picked up; when the oscillator frequency comes within a few kc/s of the standard frequency the beat note can be heard and one can then adjust the oscillator frequency until the beat frequency goes down to zero (" tuning to zero beat "). It is best if neither signal is modulated; then neither is separately audible and there is no sound to confuse the beat note. The amount of pick-up from the oscillator should be adjusted so that it is adequate but not so strong as to swamp the receiver and desensitize it for the standard signal.

Much depends on whether the frequency to be checked is already approximately known or not. If not, it is unwise to risk confusing the issue by using a superhet. Unless the signal is very weak a single-valve receiver with retroaction is satisfactory; it can be made to oscillate for preliminary searching, and then self-oscillation can be stopped when the inter-signal beat is obtained.

Precision of frequency comparison by the simple zero-beat method is limited, because there is a range of frequency each side which cannot be distinguished from zero. Using the double-beat or slow-beat modification, the precision with which two frequencies can be equalized is limited only by their steadiness; a difference of a small fraction of 1 c/s can be measured. The method is to obtain an audible beat between the standard and a third oscillation; the signal to be compared is then tuned to give the same beat frequency with this third oscillator. For example, if the standard were 1,000 kc/s, and the receiver were made to oscillate—or a separate oscillator provided—at 1,001 kc/s there would be an audible beat note of 1 kc/s. If now the signal to be compared were tuned nearly to 1,000 kc/s there would be a slow beat. One beat per second would indicate a frequency difference of one in a million. It is necessary to make sure that both the frequencies being compared are on the same side of the third frequency; in this example, 1,002 kc/s would give the same beat frequency with 1,001 kc/s as 1,000 kc/s. It is easy to settle the matter by slightly altering the third frequency; if correct, the slow beats will continue around a different audible note; if wrong, the two beat notes will diverge and destroy the slow beat.

Sometimes, as in the frequency-variation method of Q measurement (Sec. 9.29), it is desired to determine a relatively small difference between two high frequencies. If the difference is small enough to be audible, it can be measured by comparison with a calibrated a.f. oscillator. This principle is very important when the standard is an array of accurately-known frequencies, such as is produced by a multivibrator system (Sec. 6.15). Suppose a series of harmonics of 10 kc/s, locked to a standard frequency, is available. Then any frequency to be measured, not higher than the highest detectable harmonic, is within 5 kc/s of a harmonic, so there is always a beat note that is audible. E.g., if a frequency, known to be between 670 and 680 kc/s, beats with the next lower harmonic (670 kc/s) at a frequency which by comparison with the a.f. oscillator is found to be 1,565 c/s, the exact frequency is 671·565 kc/s. Even if the a.f. calibration is not better than within 5%, that would represent an uncertainty of only a little over 0·01% in this example.

For measuring frequencies that are not already known to the nearest harmonic, it is necessary to be able to identify the number of the harmonic. In the fixed-frequency type of standard, this may be done by having successions of harmonics at every 100 kc/s and every 1 Mc/s, as described in Sec. 6.15. If, however, the standard is variable, but much lower in frequency than the one to be measured (f_x), it is first adjusted so that one of its harmonics gives zero beat with f_x, the frequency of the standard then being f_1. The standard frequency is then increased until its next harmonic is heard beating with f_x. At zero beat let f_2 denote the new reading: then

$$f_x = \frac{f_2 f_1}{f_2 - f_1}.$$

With this method* there is some risk of hearing notes caused by

* "The Identification of Harmonically Related Frequencies", by L. H. Moore. *E.E.*, April 1947.

harmonics of the unknown frequency beating with higher harmonics of the standard, but these can usually be distinguished by their relative weakness, or else the error caused by using them is so large as to be obvious. To make sure, it is a good thing to check the result with another pair of standard frequencies.

10.14. CALIBRATING A R.F. SIGNAL GENERATOR

A r.f. oscillator can be calibrated from a single accurately known frequency by the use of harmonics, on the principle described in Sec. 10.11 for a.f. If the standard is a broadcast wave, its harmonics

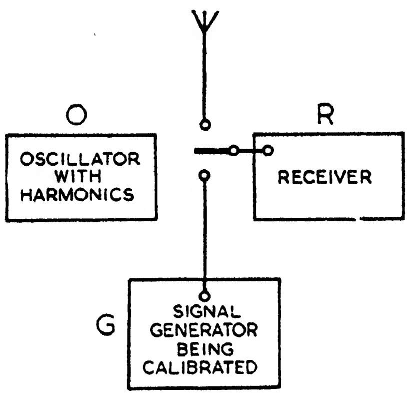


Fig. 10.12—Method of calibrating the frequency scale of a signal generator from a single known radio frequency

are not available, so it is necessary to synchronize a stable oscillator with it, as suggested in Sec. 6.14. As an example, let us suppose that the frequency scale of a signal generator is to be calibrated, using the 200-kc/s Droitwich carrier wave as the standard. Fig. 10.12 shows the essential apparatus. The main requirements of the oscillator O are that it should be precisely adjustable and constant during the time it is in use. The harmonics can be obtained by working the oscillator valve—or amplifier valve if any—so that the waveform is distorted. Stability of frequency is considered in Chapter 4.

First the Droitwich signal is tuned in on the receiver and the frequency of O set to 200 kc/s. If R can be made to oscillate, or there is an internal or external oscillator, the double-beat method can be used. Ideally, R should be left tuned in this way so as to monitor the frequency of O continuously, and a second receiver should be used for calibrating the generator G; but if this is not available the constancy of O must be reliable while R is tuned in turn to the harmonics of O (its own tuning scale should help in this); the settings of G's frequency control are noted when its fundamental gives zero beat with O's harmonics. That gives calibration points at 200, 400, 600, 800, 1,000, 1,200, etc., kc/s. Next, if closer points are needed at the low-frequency end of this range, R should be retuned to 200 kc/s (incidentally, checking that O has not drifted), and O's frequency reduced until one of its harmonics hits 200 kc/s. At the fourth harmonic, for example, there would be calibration points every 50 kc/s. Alternatively, if higher frequencies are needed, with O still at 200 kc/s R should be tuned to one of its harmonics, and O then retuned to bring its fundamental, or a lower harmonic, to that frequency. By successively shifting the frequency of O to known points and using its harmonics, a sufficient number of points can be found for plotting a curve of frequency against G's frequency-control settings.

An elaboration of this method, for enabling particular frequencies—not necessarily harmonically related to the standard—to be established, is to tune O to a frequency that differs from 200 kc/s by an amount that is measured by comparing the beat note with a calibrated a.f. oscillator. E.g., 1,770 kc/s can be established as the ninth harmonic of O when its fundamental is less than 200 kc/s by 3·33 kc/s.

A slide-rule is often useful for identifying the fundamental when a number of high-order harmonics are heard; if one harmonic comes in at 4,400 kc/s, and the next at 4,950, the moving scale is slid along until two consecutive whole numbers bridge the gap between these two numbers on the fixed scale. The only numbers that do so are 8 and 9, so these are the harmonics, and the fundamental (read below " 1 ") is 550 kc/s.

For application to v.h.f., see Sec. 12.8.

10.15. SYNTONIZING PASSIVE TUNED CIRCUITS

An absorption frequency meter (Sec. 6.16) obviously cannot be calibrated by any of the foregoing methods, which all assume the generation of a signal. If it incorporates a resonance indicator, then the method is obvious—it is tuned to resonate with a suitable signal source at a succession of known frequencies. Even if the indicator is sensitive, the signal source has to be quite powerful if the coupling is to be loose enough not to cause perceptible error. An alternative method, which does not necessitate an indicator connected to the circuit to be calibrated, is to make use of the reaction on the signal source; a milliammeter in the oscillator anode circuit gives a kick when a loosely coupled passive circuit is tuned through resonance. An oscillator with a grid-leak of rather high resistance is best for showing such an indication, which should be reduced to a minimum by making the coupling no closer than necessary.

A much more precise method of adjusting a tuned circuit to resonance is to listen to a beat note produced between the fundamental or harmonic of the oscillator to which it is coupled and some other oscillator (not coupled). Even when the coupling is much too loose to show any meter kick, the beat note can be heard to rise gradually, then fall suddenly, and then rise again, as shown in Fig. 10.13. Exact

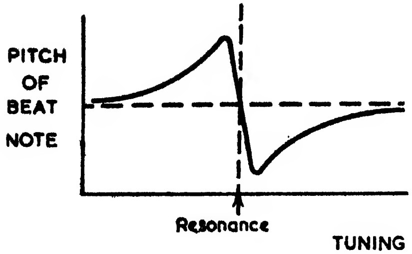


Fig. 10.13—" N curve " of the disturbing effect of a loosely-coupled circuit tuned through resonance—a phenomenon that is helpful in syntonizing circuits very precisely

resonance is at the point on the steep slope where the beat note is unaffected by removing the coupled circuit entirely or short-circuiting it. This point can be located much more precisely by using a third oscillator to produce slow beats when the frequency of the oscillator to which the tuned circuit is coupled alters even very slightly. This method, by which a non-generating circuit can be adjusted with the same order of precision as an oscillator, is described in detail by F. M. Colebrook (*W.E.*, Dec. 1931).

10.16. WOW AND FLUTTER

Undesirable fluctuations of frequency are caused in recording and reproducing systems, mainly by mechanical imperfections. The most obvious example is the disk record with a hole that is not exactly in the centre, making the pitch of the sound deviate at the frequency of the turntable revolutions. Frequency modulation of this kind is called wow if it occurs at frequencies between 0·1 and 20 c/s, and flutter if it is 20 to 200 c/s. A common cause of wow is eccentricity of the capstan in tape recorders. Variations below 0·1 c/s are called drift, and are not likely to be noticeable unless unusually large—except perhaps by listeners gifted with absolute pitch. Frequency variations of these kinds are almost inevitably accompanied by amplitude variations, but these usually cause only a minor part of the unpleasantness, so it is sufficient to measure the frequency modulation. (It should be mentioned, however, that amplitude modulation alone may be produced in a.f. systems, especially by 50 c/s from the mains or by sub-audible oscillation in amplifiers, with extremely unpleasant results.)

The unpleasantness of wow (using that term now to include flutter) depends of course on the depth of modulation, expressed as a percentage of the frequency modulated. It also depends on the frequency with which it occurs, and to some extent on its waveform. And to a very large extent on the kind of programme affected; wow which is quite imperceptible on speech may be intolerable on piano music. The relationship between percentage wow and the subjective effects was investigated in detail by the B.B.C. engineers, A. Stott and P. E. Axon (*Proc. I.E.E.*, Part B, Sept. 1955). Their results show that wow not exceeding 0·1 % is acceptable, but 0·5 % is not. Incidentally, it is not always clear whether percentages refer to peak, peak-to-peak, or r.m.s. values. E.g., fluctuation from 2,970 to 3,030 c/s could be 1%, 2% or 0·7%. Where not stated, r.m.s. is assumed (B.S. 1998: 1953).

Simulation of wow cannot readily be done electronically, because (unlike the action of frequency changers) it is the same percentage of all frequencies in the programme; not the same frequency shift for all.

Measurement is usually done on a " programme " consisting of a pure tone, the standard frequency for which is 3 kc/s (B.S. 1998: 1953). The apparatus is virtually an f.m. receiver, modified considerably because of the low frequency of the carrier and the smallness of the deviation. A method is described, with details of circuitry by O. E. Dzierzynski (*W.W.*, Nov. 1955). Equipment is also available commercially from Kalee G.B. Ltd.

10.17. MAGNETIC FLUX

In view of the scale on which magnets are used in the radio industry—in loudspeakers, gramophone pickups, microphones, television focusing, and microwave generators, for example—some reference to practical magnetic measurements should not be omitted from even a book like this. An account of the established methods and equipment is given in F. G. Spreadbury's *Permanent Magnets*, Ch. VII (Pitman).

The classical instrument for measuring flux is the Grassot fluxmeter. Although portable, it is fairly delicate and not very cheap, so the alternative described by P. L. Taylor (*W.W.*, April 1951), which is no more than a special type of valve voltmeter, readily made from parts likely to be found in the laboratory, is interesting. The first stage is a conventional coupled cathode balanced pair of EF37 valves, by which an amplification of the voltage induced in a search coil by moving it out of the field to be measured is set up between the anodes. The coupling to the output stage—a pair of SP61 valves with a 0 to 1 mA meter between the anodes—is a simple *RC* circuit that integrates the voltage with respect to time, so that the output reading is proportional to the change of flux linking the search coil, irrespective of the time taken to make the change, provided it is reasonably small. If the search coil can be made to embrace the total flux in question, then that is measured directly. Flux density can be measured by using a search coil small enough in turn area for the linking flux to be uniform, and dividing the flux by that area.

Modern fluxmeters usually make use of the Hall effect in a small strip of semiconductor mounted at the end of a probe, with which quite narrow magnetic gaps can be explored (Fig. 10.14). This type has the great advantage of giving a continuous direct reading of flux density. A battery in the main part of the instrument passes a known current lengthwise through the strip, and a millivoltmeter measures the difference of potential between two contacts at the sides of the strip, as in Fig. 10.15. In the absence of a magnetic field there is of course no p.d. between them, but a field passing through the strip (at right angles both to the current and to a line drawn between the contacts) pushes,

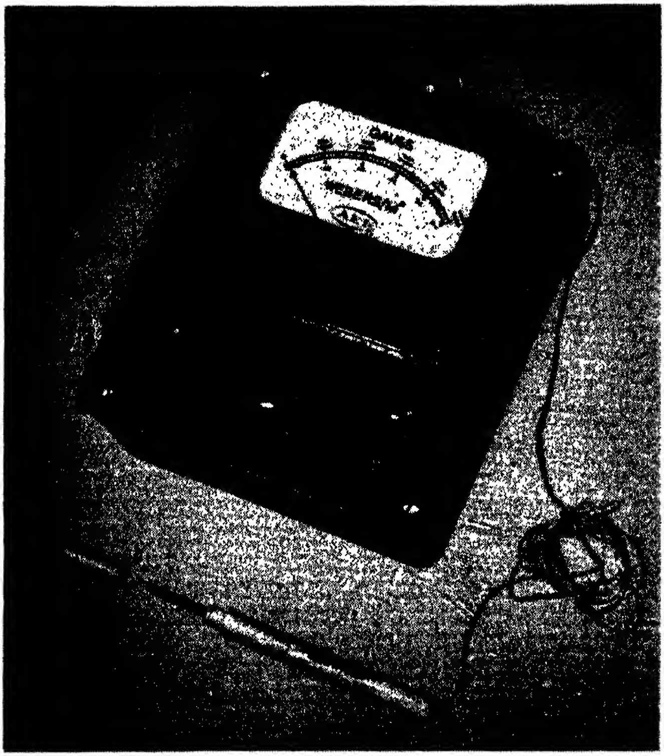


Fig. 10.14—A.E.I. direct-reading flux-meter, employing the Hall effect in germanium. The point of the probe is held in the magnetic field to be measured. (Associated Electrical Industries Ltd.)

as it were, the current to one side, producing a p.d. directly proportional to the current and to the field.* The millivoltmeter can therefore be directly calibrated in flux density.

Hall effect is very small in metals, owing to the slow onward speed of electrons constituting the test current; it is much more easily measurable in germanium, and better still in certain semiconductor compounds, notably indium antimonide.

The A.E.I. instrument shown, which uses a germanium probe, has three ranges, with full-scale readings 1, 5 and 25 kilogauss; in m.k.s. units, 0·1, 0·5 and 2·5 Wb/m^2.

One other and quite different method should be mentioned. The fact that it is based on somewhat esoteric phenomena in atomic physics may scare off some of the "practical" workers, but the more adventurous can hardly fail to be excited by a visual demonstration of resonances in the component parts of atoms. To reassure the timid,

* For a fuller explanation see "Hall and Holes", *W.W.*, Dec. 1958.

one can point out that ordinary electric currents depend on atomic phenomena which are subjects for most abstruse scientific theories, but that does not prevent infants from using them quite successfully.

Skipping the theory, one can say that certain paramagnetic substances, such as diphenyl picryl hydrazyl, are equivalent to absorption wavemeters having the exceptional feature that their frequencies of resonance depend not on the substance but are proportional to the

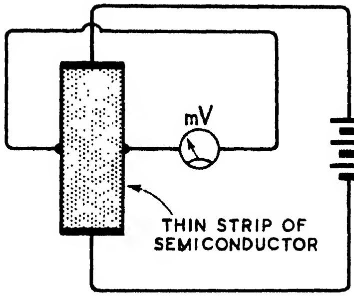


Fig. 10.15—Essentials of magnetic-field indicator making use of Hall effect

strength of the magnetic field in which it is placed. Such a frequency is found by noting the absorption from an oscillator connected to a coil wound around a sample of the substance in the magnetic field to be measured and tuned through the frequency somewhat as in Sec. 10.15.

One absorption frequency is caused by proton resonance:

$$f_p = 42{\cdot}6B \qquad [\text{Mc/s; Wb/m}^2]$$

Another is due to electron spin resonance:

$$f_e = 28{,}000B \qquad [\text{Mc/s; Wb/m}^2]$$

The latter can be detected in quite weak magnetic fields, such as 0·001 Wb/m² (10 gauss), corresponding to 28 Mc/s.

The modern theory of particle magnetism and resonance is given at length in *Electricity and Magnetism* by Bleaney and Bleaney (Clarendon Press, 1957). " Paramagnetism " (*W.W.*, July/Aug. 1959) is a simple introduction. A more comprehensive treatment with particular reference to proton resonance and measurement of magnetic fields was given by " Quantum " (*E. & R.E.*, June 1957)*. G. B. Clayton (*W.W.*, Feb. 1960)* has given an outline of theory and details of actual apparatus, supplemented by A. G. A. Rae (*W.W.*, July 1960, p. 347).

* In these articles some confusion occurs in the use of the term " gyromagnetic ratio ", for which read " magnetogyric ratio ". Note that

$$\gamma = \text{magnetogyric ratio} = \frac{\text{magnetic moment of particle}}{\text{angular momentum of particle}}$$

$$= \frac{1}{\text{gyromagnetic ratio}} = 2\pi \times \text{the constant connecting } f_p \text{ and } f_e$$

with B in the above equations.

CHAPTER 11

Measurement of Equipment Characteristics

The subject matter of this chapter is almost unlimited, and no more can be included than a selection of the more important and frequently required measurements. Information on instruments and methods mentioned is given in earlier chapters. For information on fault-finding and adjustment of equipment, see Cocking's *Wireless Servicing Manual* (Iliffe).

11.1. AERIAL IMPEDANCE

Although the capacitance, inductance and resistance of an aerial-earth system are distributed throughout itself, like any other impedance it is equivalent *at any one frequency* to a resistance and a reactance, in series or in parallel as one may prefer (Sec. 14.12). So it can be measured by the substitution methods described in Secs. 9.25 to 9.33, except that owing to the rapid change of impedance with frequency the frequency-variation method should not be used.

The tuned circuit used for the measurement is normally connected to earth at one end, so all that has to be done is to connect the aerial to the other and note the change of capacitance and parallel resistance, and, if desired, convert to equivalent series values. Most ordinary outdoor broadcast receiving aerials are equivalent (at low and medium r.f.) to rather " lossy " capacitors of the order of 200 pF. But if connected, as some are, through a r.f. transformer and feeder, the impedance at the feeder terminals may be too low to measure by substitution across a tuned circuit and may have to be connected in series, as explained in the sections just mentioned. Since one terminal will be earthed, it is necessary to connect to the earthy side of the measuring apparatus.

For measurement purposes, frame and ferrite aerials can be regarded as inductors.

11.2. TRANSMISSION LINES OR R.F. CABLES

A coaxial or parallel-wire line is like an aerial in that its parameters are distributed, but because its radiation resistance is deliberately made small its Q is much higher, and owing to its regular geometrical form its impedance varies in a more regular manner with frequency. When in use it should be matched by connecting it at both ends to impedances (ideally, resistances) of a certain value called the characteristic impedance (Z_0), such that the impedance measured from either end does not vary with frequency but is equal to Z_0. This Z_0 can be regarded as the value of load impedance that absorbs the whole of the power sent along the line; with any other value, part of the power is reflected back to the source. Going to extremes, if the far end is either open- or short-circuited the line behaves as a resonant circuit of high Q,

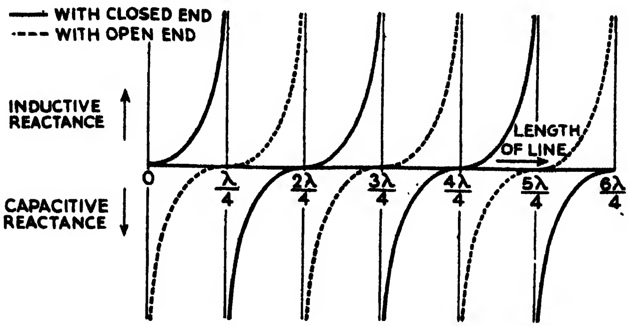


Fig. 11.1—Showing how the reactance of short-circuited and open-circuited lines varies with length

alternately series- and parallel-resonant as frequency is varied. At intermediate frequencies it is equivalent to a positive or negative reactance. A similar cyclical sequence of resistances and reactances is obtained at a fixed frequency if the length of the line is varied. Fig. 11.1 shows how the reactance of a resonant line varies with its length, expressed in units of wavelength corresponding to the frequency employed. Where reactance is zero, the impedance is a low resistance (ideally zero) and where it is infinite the impedance is a high resistance (ideally infinite). Short lengths of line are often used as efficient tuning circuits or reactors, especially at frequencies so high that lumped circuits are unsatisfactory.

The resistance and reactance of a length of open or shorted line at a specified frequency can be measured by any of the methods suitable for an impedance of that order of magnitude at that frequency. As the frequency is often high, this may present some difficulty (Sec. 12.10), but if the dimensions are regular the impedance can be calculated (Sec. 14.27).

Z_0, too, can often be more easily calculated than measured, at least if the conductors are air-spaced. In that case the attenuation or loss of signal is small, and its velocity along the line (*phase velocity*) is very nearly as great as through space. But cables having solid dielectric are a different matter. The phase velocity may be less than half that in space and the loss per 100 ft is often several dB. As Fig. 11.1 shows, there are points (midway between the vertical lines) where the reactance of a line with the far end open is equal and opposite to the reactance of the same line at the same frequency with the end closed. It is also equal in magnitude to Z_0, so one method of measuring Z_0 is to connect the line for measuring its capacitance by some substitution method suited to the desired frequency, and then adjust either the length of the sample or the frequency so that the capacitance with the end shorted (C_s) is equal to minus the same capacitance with the end open (C_p). Actually there is no need to get them exactly equal; if they are approximately equal there is negligible error in assuming that

$$Z_0 = 1/2\pi f\sqrt{-C_sC_p}$$

As a guide, at frequencies in British television Band I and Z_0 about 75 Ω, the capacitances to be measured are of the order of 40 pF. Since Z_0 changes little with frequency, the exact frequency is usually unimportant and a suitable measuring-point can be obtained by frequency adjustment. The longer the cable the less the frequency shift to cover the whole gamut of line capacitance.

Loss can be measured by adjusting the frequency to bring the sample (either shorted or open at the far end) to " parallel " resonance, at which it behaves like a high resistance R_x, which can be measured as for dynamic resistance. Then the loss is

$$\begin{aligned} \alpha &= Z_0/R_x \text{ nepers,} \\ &= 8{\cdot}686\, Z_0/R_x \text{ dB} \qquad \text{(Sec. 14.34)} \end{aligned}$$

An alternative method is to measure R_x as above, and then reverse the condition at the far end and again measure the resistance, which is now a low value (r_x) corresponding to " series " resonance:

$$Z_0 = \sqrt{r_xR_x}$$

and the loss is

$$\sqrt{(r_x/R_x)} \text{ nepers}$$
$$\text{or } 8{\cdot}686\sqrt{(r_x/R_x)} \text{ dB.}$$

A disadvantage of the second method is that r_x is too low to be measured with the same substitution connection; it has to be measured in series.

The phase velocity is

$$v = lf/n$$

where l is the length of the line and n the number of wavelengths therein. These wavelengths, marked in Fig. 11.1, are not what is usually understood as " the " wavelength, equal to c/f, where c is the velocity of waves through space; they are wavelengths along the line, as would be indicated by twice the distance between successive maxima or minima of standing waves. These can be detected on parallel-wire lines by a suitable indicator, but on coaxial cables n has to be deduced by varying the frequency. Fig. 11.1 shows that a maximum input resistance occurs whenever the length of a shorted line is an odd number of quarter-wavelengths. It follows that if one such maximum occurs at frequency f_1 and the nearest higher frequency for another is f_2, and v is assumed to be the same at both,

$$n = f_1/2(f_2 - f_1)$$

and
$$v = 2l\,(f_2 - f_1)$$

The parameter β is often used, the phase shift per unit length:

$$\beta = \frac{2\pi}{\lambda} = \frac{2\pi f}{v}$$

The method for Z_0 and phase velocity described by L. B. D'Alton (*E.E.*, Jan. 1958) is worth noting because the apparatus required comprises only a signal generator, radio receiver with metered output,

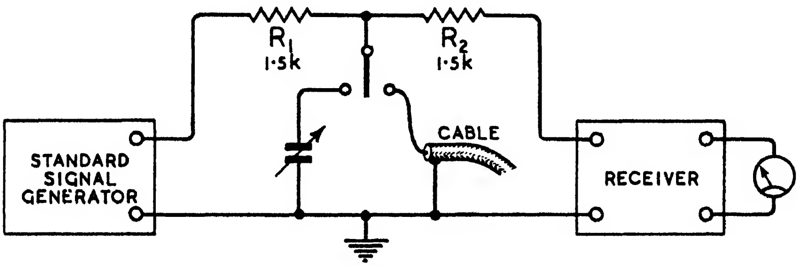


Fig. 11.2—Easily assembled apparatus for measuring characteristic impedance and phase velocity of a sample of cable

and calibrated variable capacitor. An open-circuited sample of cable is used, preferably about 2 metres long. Provided this is less than $\lambda/4$, and loss is negligible, it is equivalent to a capacitance C, measured by simple comparison as shown in Fig. 11.2, which is self-explanatory. R_1 is for increasing the effect of a small change in capacitance on the meter reading, and R_2 as a buffer between the capacitance and the receiver. Z_0 is derived from

$$Z_0 = \frac{\tan \dfrac{2\pi f l}{v}}{2\pi f C}$$

the symbols having the same meanings as before. For best accuracy f should be about $23/l$ Mc/s. If v is not known, repeat the measurement at double the frequency, obtaining a new reading C'. Then

$$\tan \frac{2\pi f l}{v} \left(= \tan \beta l \right) = \sqrt{1 - \frac{C}{C'}}$$

A method for computing Z_0, v and α from readings taken on a half-wave sample of cable with a Q meter is fully described by J. Shekel (*E.E.*, Dec. 1954).

A useful series on cable characteristics and their measurement appeared in *G.R.E.*, May–Aug. 1957. The graphs of Z_0 and v against frequency are level except for a " step " caused by the onset of skin effect.

11.3. A.F. TRANSFORMERS

The effective frequency band of an a.f. transformer is limited at the low end by shunt inductance and at the high end by leakage or series inductance. Shunt inductance, so called because it acts as a by-pass to the transformer load impedance, is simply the inductance of the primary. It should be measured at a low frequency, to ensure that the effect of stray capacitance is negligible; 50 c/s is usually suitable. Working conditions should be reproduced as nearly as possible during measurement, as regards amplitude of a.c. and d.c. (if any). Suitable methods are described in Secs. 9.18 to 9.22.

Leakage inductance is relatively small, and is a measure of the extent to which the primary is not completely coupled to the secondary. The equivalent total leakage inductance referred to the primary is the inductance measured on the primary side when the secondary is short-circuited. Although it affects the high frequency end, there is no reason why it should not be measured at the same low frequency as the shunt inductance, except perhaps that its reactance is then rather small.

Self-capacitance is an important parameter of intervalve transformers, but these are now rarely used; in output transformers it often plays only a minor part. But it may have to be considered when the output transformer is in a negative-feedback loop, especially since it is increased by the means adopted in the better transformers to reduce leakage inductance—the interleaving of primary and secondary windings. With this arrangement one may not be safe in assuming that the stray capacitance is equivalent to a constant capacitance in shunt with the primary and independent of frequency, and instead of trying to measure it as such it is better to measure the phase shift (Sec. 10.7).

The efficiency of an output transformer can be measured by means of an output meter and a.f. signal source. The source should be adjusted to give a suitable amount of power, measured first by the output meter connected directly to it, and then through the transformer. If the transformer ratio is $n : 1$ and its optimum load resistance R, the resistance of the output meter should be as near as possible to n^2R for the first reading and R for the second. The efficiency is the ratio of the second reading to the first, and should be measured at the lowest, highest, and one or more intermediate frequencies. Alternatively this power ratio can be expressed in dB (Table 14.17) and is then called the transformer loss. For this method it is necessary for the output-meter transformer not only to have the appropriate pair of ratios but also reasonably equal efficiencies at these ratios, at all frequencies concerned. If there is any doubt about this—or if an output power meter is lacking—the measurement can be made by a valve voltmeter or even c.r.o. across the appropriate resistances.

An important characteristic is non-linearity, reckoned as the amount of distortion caused under specified conditions by a specified signal amplitude, or alternatively the maximum signal power for a specified amount of distortion. Perhaps the most useful information, covering both of these, is a curve of distortion against amplitude or power. Methods are the same as for a.f. amplifiers; see Secs. 11.12 to 11.16.

It is of course most essential that working conditions or their equivalents are reproduced—the same amount of d.c. and the same impedances connected to the windings.

Ref: "Harmonic Distortion in A.F. Transformers", by N. Partridge. *W.E.*, Sept. to Nov. 1942.

11.4. LOUDSPEAKERS

For many purposes a loudspeaker and its transformer can be considered as a single unit, and measurements of its impedance referred to the primary of the transformer. Theoretically, with a perfect transformer of ratio $n : 1$ the impedance regarded from the primary side is n^2 times the impedance of the speaker itself, but in practice there are slight divergences due to transformer characteristics, such as leakage inductance; and often there is considerable power loss in the transformer.

A distinction must be drawn between the purely electrical impedance of the speaker and the working impedance, which includes what is known as the motional impedance, due to the back-voltage induced by the movement of the coil in the magnetic field. The difference is hardly noticeable in the upper parts of the frequency scale, but at the mechanical resonance of the cone and coil the motional impedance is the main part. It can be determined by measuring the total impedance firstly with the coil free and then with it firmly clamped in the normal position in the gap, and taking the vectorial difference (Sec. 14.12). The methods of measurement are exactly the same as those described in Secs. 9.18 to 9.22 for reactance and resistance of coils; particularly those with iron cores. D. E. L. Shorter has shown (*W.W.*, Nov. 1950) how a simple resistance bridge with a valve voltmeter as detector can be made to give readings directly proportional to coil velocity. These can be used to investigate the performance of the loadspeaker and its cabinet below 500 c/s, which is especially important, since it includes the main resonance, commonly in the region of 80 c/s, and the greatest amplitude of movement.

The severity of the resonance itself; the amplitude of coil movement at which it goes outside the uniform magnetic field, causing modulation of the upper frequencies simultaneously present; the possibility of mechanical rattle at the large amplitudes associated with low frequencies; and the production of sub-harmonic tones due to flexing of the cone at certain frequencies—all these can be investigated by an oscillator capable of giving a practically undistorted output of several watts. The mechanical movements of the coil and cone can be seen in slow motion by examining it by the light of a neon lamp fed from another oscillator differing from half the frequency of the first by only about 1 c/s, or alternatively by an oscillator working at nearly the same frequency and with d.c. superimposed so as to provide a pulsating instead of an alternating supply.

The modulation effect mentioned above can be investigated by the same methods as for amplifiers (Sec. 11.16), the only difference being the need for a linear microphone to translate the acoustical output into electrical form.

The acoustic characteristics of a loudspeaker are much more important than the electrical ones, but unfortunately much more difficult to measure. Not only does one need a calibrated high-quality microphone covering the whole a.f. band; the tests cannot be carried out in an ordinary room, because they would be vitiated by its complex resonance pattern. An anechoic chamber is required, which is one lined to a depth of several feet with special acoustic materials arranged in irregular wedge formations, and if properly constructed is very costly. Open-air testing suggests itself, but by the time a large field has been purchased, on a site far from such noise sources as road, rail or air transport, and towers erected for supporting loudspeaker and microphone 30–40 ft above the ground, it is not likely to show much saving and is even less convenient. So " white noise " tests are very attractive. A noise generator can easily be made at low cost (e.g., Fig. 4.49), and produces a signal whose power (or the square of its voltage) between any two frequencies is directly proportional to the frequency difference. The output from the loudspeaker under test, fed from this signal (with or without its own amplifier) is modified by the frequency characteristic of the speaker and amplifier. So if a set of band-pass filters, each selecting the same bandwidth at different frequencies in the a.f. band, is used in the amplifier following the microphone, the output readings therefrom indicate the relative loudspeaker response at those frequencies. In practice the filters are more likely to select an octave or other frequency *ratio*, in which case the readings would have to be divided by the mean frequency of the band. The subject is dealt with by J. Moir in *S.R. & R.*, Feb. 1959.

The following are some references to information, especially on anechoic tests:

B.S. 2498 (*Recommendations for Ascertaining and Expressing the Performance of Loudspeakers by Objective Measurements*).
Acoustic Measurements, by L. L. Beranek (Chapman & Hall, 1949).
See also " Loudspeaker Response Curves ", *W.W.*, Feb. 1952, for a method of recording frequency response curves so as to show non-linearity.

11.5. INPUT/OUTPUT MEASUREMENTS: STANDARD TERMS

Most of the remainder of this chapter deals with receivers and parts thereof such as amplifiers, for which the most usual type of measurement is that shown in broadest outline in Fig. 3.1*b*. Certain terms have been standardized (B.S.2065 : 1954) and these are shown in Fig. 11.3, where the previous outline is filled, though still only generally. The circuits shown inside the " boxes " are of course not the actual circuits of any particular apparatus, but only their simplest electrical equivalents as viewed from the terminals (Sec. 14.21).

Where the full meaning is clear from the context, " input " is sometimes used to mean input terminals, input impedance, input voltage, or input power; and similarly for " output ", etc.

Except where otherwise stated, it is generally assumed that the source and load impedances are equal to those for which the item under test was designed. To make this so, various matching devices

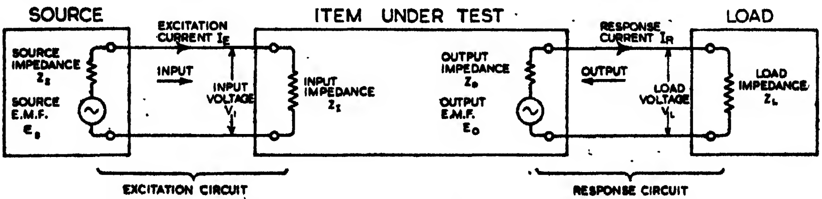


Fig. 11.3—Diagram showing standard terms and symbols used in connection with tests on amplifiers, receivers, etc.

such as transformers and dummy aerials are commonly used. The greatest power is received from a given output when (1) the load resistance is equal to the output resistance and (2) any output reactance is neutralized by equal load reactance of opposite sign; this amount of power is called the *available power*, and is equal to $E_0^2/4R_0$. Similarly for the source. The designed load is not always the load that would receive maximum power; where the load is a loudspeaker it is usually more important to design for minimum distortion, for the output voltage can easily be increased to make good any power loss. And in a.f. amplifiers the input impedance may be practically infinite, for voltage rather than power is the main consideration.

The amplification or gain of an amplifier is the ratio of load power (or voltage) to input power (or voltage). Strictly, " gain " refers to power ratio. If there is any room for doubt, the basis of reckoning should be specified—e.g., " voltage amplification ", or " power gain ". Power ratios are commonly expressed in decibels, as explained in Sec. 14.34. Only one thing need perhaps be re-emphasized here—that a voltage (or current) ratio cannot correctly be expressed in dB unless account is taken of the impedances across which the voltages exist (or through which the currents flow). For the matter of that, a voltage ratio itself is not very informative unless the impedances are specified.

11.6. A.F. AMPLIFIERS: GAIN

The obvious way to measure gain is to apply a suitable signal to the input and observe its strength there and at the output. There are, however, diverse methods of observing. The best, probably, is to insert a calibrated attenuator and adjust it until the voltages at input and output of the combination are equal. The gain due to the amplifier

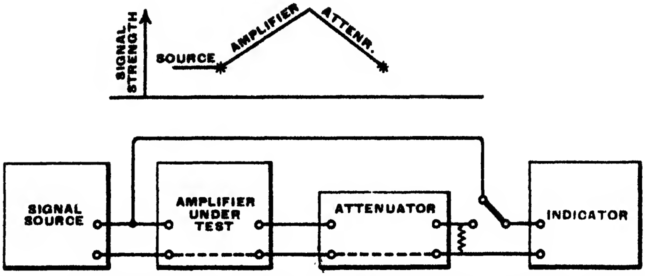


Fig. 11.4—Block diagram of the arrangement of apparatus when measuring the gain of any amplifier by comparing it with the loss introduced by a calibrated attenuator. If the attenuator is continuously variable over the range concerned, the only purpose of the meter is to adjust the signal level to equality at the two points shown by the stars in the signal-strength diagram; no meter calibration is needed. Note that both source and meter are at relatively low signal levels

is then equal to the loss in the attenuator. This makes it easy to use one indicator instead of two for observing the signal levels; moreover, it need be only a simple indicator, for it has to do no more than enable two signal voltages to be equalized, which is easier and more accurate than measuring the ratio between two that are likely to be of quite different orders of magnitude.

The attenuator can be inserted either ahead of, or following, the amplifier. Fig. 11.4 shows the latter arrangement. The terminations of the units must be considered: the input impedance of the attenuator should be appropriate to the output of the amplifier, and the output of the attenuator should be properly terminated. In this respect the alternative order—attenuator first—is usually more convenient; the amplifier can be used with its normal load, such as a loudspeaker, and the amplifier input impedance may well be high enough for a simple potential-divider type of attenuator to be used (Sec. 6.19). Incidentally, although the use of negative feedback may greatly reduce the output impedance of an amplifier, its optimum load impedance is hardly affected thereby.

The order of connecting also affects the types of signal source and indicator. As the upper part of Fig. 11.4 shows, the amplifier-first order results in low level of signal voltage at the indicator, and would be suited to a low-output source, but not to an insensitive indicator such as an unamplified cathode-ray tube. With the amplifier and attenuator interchanged, the signal level reaching the amplifier from a weak source might not be clear of noise, hum, etc., but would be satisfactory if the source output were comparable with the normal output of the amplifier.

A single indicator can be switched from end to end, as shown, in which case it need not be calibrated, nor even be particularly free from drift or frequency error, provided it does not take enough current to alter the signal when it is switched in. A valve voltmeter, even if only extemporized, is an obvious type. A metal-rectifier meter, or even a pair of headphones, may be used if their impedance is high enough relative to the signal circuits. It is merely a matter of adjusting to equality, not of actual measurement. It need hardly be mentioned that a sensitive valve voltmeter must not be switched in such a manner as to open-circuit its grid between the two positions, and that precautions must be taken to exclude non-signal potentials such as h.t. voltages. Both objects are achieved by using a grid capacitor and leak, but even then there is a possibility of momentary violent surges on switching over, which may injure a sensitive instrument unless it is temporarily desensitized.

The simple switching shown assumes that the earthy side of each instrument is common; if not, a double-pole switch is of course needed.

The alternative is to use two indicators and no switching, so that their impedance does not matter so much; but it is necessary to check their readings one against the other at all frequencies if the actual gain is to be measured accurately. Neither meter need be calibrated, unless the signal level is to be measured at the same time. At least some

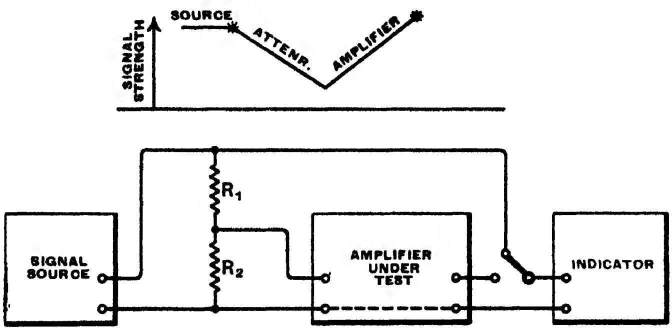


Fig. 11.5—It may be better in some cases for the amplifier and attenuator to be at low signal levels, source and meter being high. This is done by putting the attenuator ahead. This diagram also shows the use of a simple fixed or semi-variable attenuator to bring the meter points to approximate equality. A meter calibration is then needed as well as the attenuator calibration, but the need for wide range is avoided

idea of the signal level ought to be known in order to avoid overloading the amplifier or perhaps even the attenuator.

There are of course other variations of the method. If the indicator is confined strictly to the job of showing equality, the attenuator has to be continuously variable. A step-adjusted attenuator can be used, however, by setting it to give the nearest approach to signal equality, and interpolating with the indicator. For example, if the attenuator reads 25 dB and the ratio of output to input overall is 1·15 (that is, 1·2 dB) the gain is 26·2 dB.

In fact, lacking any proper attenuator, one can use a simple potential-divider (R_1R_2 in Fig. 11.5) to attenuate the signal by a known amount, bringing the signals to be indicated at least to the same order of magnitude. There is then less risk of feedback due to switching, and the input signal can be read on the same meter range as the output. This, incidentally, is an example of having the attenuator first.

Such a step-down is almost certain to be needed if the indicator is a c.r. tube. R_1 can then be a fairly high fixed resistance and R_2 a low variable. In one method the tube is used simply as a kind of valve voltmeter, one pair of plates being out of use and the other switched alternately to input and output; the attenuator is adjusted until the deflections are the same. The time base can be switched on to the idle plates for checking the signal waveform. In the other method the input is connected to one pair and the output to the other. The result is a trace of the type shown in Fig. 10.8*a*. This is helpful for showing phase difference between input and output, or—as in Sec. 11.12—for indicating distortion, but only at 0° and 180° is it convenient for indicating the relative magnitudes, by the slope of the line. If the deflection sensitivities of the pairs of plates were the same, equality would be shown by a 45° slope, but as in general they are different it is necessary to allow for this by first connecting the one signal to both pairs and marking the slope of the line on the face of the tube.

For measuring separate stages of an amplifier there is a risk of working conditions being upset by the connecting of the indicator at intermediate points, and it is better to adapt the method described for r.f. amplifiers in Sec. 11.22 (Fig. 11.20), in which an indicator is kept connected at the output and the source is connected through an attenuator in turn to various input points. If at any of these the signal is liable to be appreciably short-circuited by the output impedance of the previous stage, this impedance should be open-circuited while the reading is being taken.

11.7. ATTENUATION OR LOSS

If the unit to be measured introduces a loss, the attenuator with which it is compared must be connected in parallel with it, instead of in cascade; otherwise the procedure is the same. Fig. 11.6 illustrates the connections when the known attenuator is a simple potential-divider. For example, suppose the voltmeter reading in position 1 is 1·3 and in position 2 is 1·5, and that R_1 and R_2 are 900 Ω and 100 Ω respectively. Then, assuming the load is also 100 Ω, the " gain " is $\frac{1\cdot5}{1\cdot3} \times \frac{100}{900 + 100}$ or 0·115, which is $-$ 18·8 dB.

11.8. FREQUENCY CHARACTERISTICS

Obtaining the frequency characteristic of an amplifier, filter, etc., is just an extension of the procedure already described, the gain or loss being measured at a sufficient number of frequencies to provide data for a frequency curve. This is plotted on curve sheets having a logarithmic horizontal scale for frequency and a linear vertical scale for dB. A standard kind has three decades of frequency, which can be marked

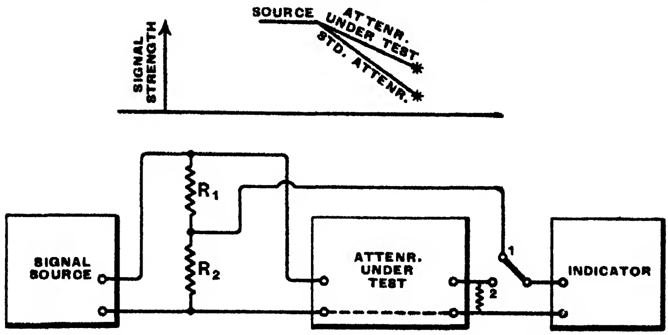


Fig. 11.6—Here two attenuators are being compared, and the only difference in method is that they are connected in parallel instead of series. If possible the standard is adjusted to make the two stars coincide

to run say from 20 c/s to 20 kc/s. If the frequency of the signal source is continuously variable, it should be swept through the whole range while an eye is kept on the output indicator to make sure that no significant feature escapes unplotted.

It is when taking a frequency characteristic that one realizes the immense advantage of having a constant-output source. If the input can be relied upon to be constant at all frequencies, all that one has to do on shifting to a new frequency is to adjust the attenuator to keep the output meter reading constant and then to read the attenuator. Switching and input adjustment are avoided. Care must be taken, however, that overloading does not happen at any frequency.

If much work is to be done on the testing and design of a.f. equipment, the continual plotting of frequency curves becomes very tedious. It may pay to have special apparatus for tracing curves automatically or semi-automatically. Various methods have been devised, some depending on mechanical operation of the oscillator frequency control linked with the frequency base of the recorder or oscilloscope, and some entirely electrical in action. It is important that the frequency sweep is not *too* rapid for the apparatus under test to follow. The firm of Bruel & Kjaer specializes in equipment of this type, which is provided also by General Radio (see *G.R.E.*, June 1959). The use of " white noise " for measuring frequency characteristics has already been mentioned in Sec. 11.4.

11.9. AVOIDANCE OF OVERLOADING

In the foregoing descriptions of gain and frequency characteristic measurements, the only references to signal amplitude have been warnings to keep it well within the handling capacity of the apparatus. Up to a point, the output of a normal amplifier is almost exactly proportional to the input—in other words, the amplifier has a linear amplitude characteristic—but beyond that point the characteristic bends over in some such way as is shown in Fig. 11.7. It is evident that if curve 1 is the characteristic of an amplifier it does not matter what input is used to measure its gain so long as it does not exceed A′ (or the output exceed A″). The need for caution occurs when the gain is measured at other frequencies or conditions, because if at some other frequency the gain is higher, as shown by curve 2, the original maximum allowable input A′ might go beyond the new

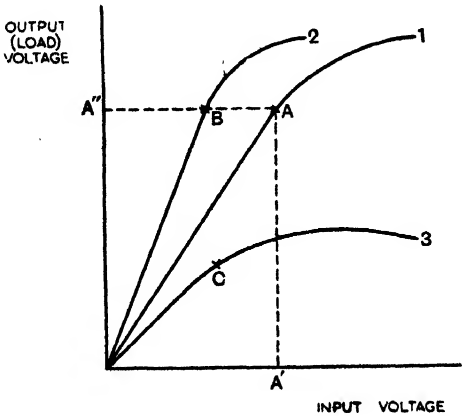


Fig. 11.7—Showing how signal level that is within the capabilities of an amplifier in one set of conditions may overload it in another

overload point B. It happens that a constant output would have avoided this error, but it is possible for the characteristic to change to curve 3, when neither input nor output voltages previously certified as safe could be allowed.

11.10. DETERMINATION OF OVERLOAD POINT

The position of the overload point is not absolutely definite, because if an apparently straight portion of characteristic is accurately plotted on a sufficiently large scale it is found to be not quite straight. Any curvature at all is evidence of some distortion, and it is just a question of how much is allowable. It is desirable to be able to give an actual figure, so that exact comparisons can be made. However, if the apparatus for doing this is lacking, one can judge fairly well, after a little experience, how much visible curvature of an input/output graph represents appreciable audible distortion. A gain-measuring set-up such as Fig. 11.5 is used, and the excitation increased until the gain shows definite signs of change. A much quicker method, suitable for production tests, is to use a pair of ganged attenuators, one ahead of the amplifier and the other behind, designed to keep the total attenuation the same at all settings. So long as the amplifier is linear, therefore, the final output is constant at all attenuator settings, but beyond the overload point it changes. A specified change can be chosen as marking the overload point and if the source is always the same the attenuator setting at that point is a measure of the power-handling capacity of the amplifier.

An input/output curve can be plotted in terms of voltage or power but not a mixture of both. If the input is measured in volts, the readings should be squared if they are to be plotted against output readings in watts, as in Fig. 11.8. This curve appears to be quite linear only as far as 1, but that is a rather conservative overload point. Reasonably good quality may be expected with the very moderate curvature up to 2, but an output corresponding to the third point would probably be marred by very noticeable distortion. Use of negative feedback in amplifiers tends to straighten the lower or working part of the curve and accentuate the overload point.

Instruments for measuring output power are discussed in Sec. 5.7;

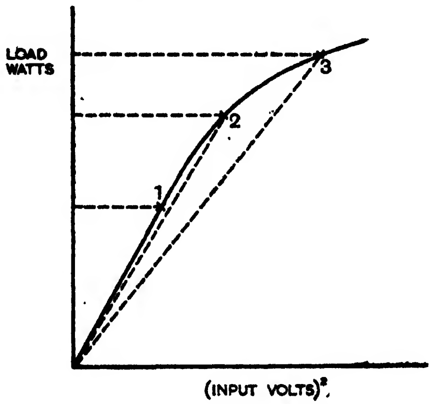


Fig. 11.8—Typical input/output characteristic of an amplifier or other unit, showing how estimation of the overload point is a matter for judgment

see also P. E. Dyson (*M.I.*, March 1957). The usual methods involve measuring the voltage across a known resistance—the load resistance. If the voltage is measured across a loudspeaker or similar load, it must be remembered that the load impedance is neither constant in magnitude at all frequencies nor non-reactive, and the nominal resistance of the loudspeaker multiplied by the square of the voltage may differ considerably from the actual watts.

11.11. LOAD-RESISTANCE CHARACTERISTICS

It is often instructive to take a load-resistance characteristic, which is a curve of load watts against load resistance. There are two varieties, which must not be confused. In each case the load resistance is varied and the corresponding watts in it observed, but in one the excitation is kept constant at some level comfortably below the overload point, and in the other it is adjusted for each load resistance until a specified limit of distortion is reached. In the first case the maximum power is obtained when the load resistance equals the effective output resistance of the valve feeding it (and is a convenient method of measuring that quantity); in the other, the load resistance receiving the greatest *useful* power, and the amount of that power, are found; the optimum resistance depends on a number of conditions, but is usually at least twice a triode-valve resistance and only a fraction of a pentode resistance. Note the *effective* output resistance is mentioned above; if negative feedback is used, such resistance may be quite different from the actual valve resistance.

11.12. OBSERVATION OF NON-LINEARITY BY C.R. TUBE

Not only is estimating non-linearity by looking at a curve of the type shown in Fig. 11.8 rather indefinite; the curve is unsuitable as a basis for more precise analysis of non-linearity. The points on it do

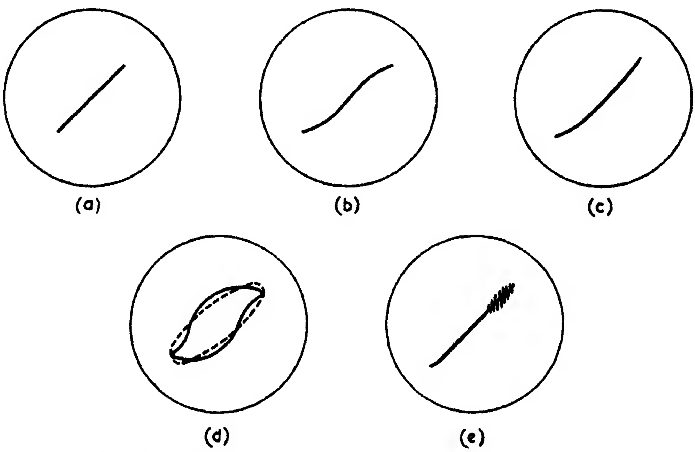


Fig. 11.9—Representative c.r. tube traces when input signal is applied to X plates and output to Y, showing the appearance of (b *and* c) *non-linearity,* (d) *phase shift and* (e) *parasitic oscillation*

not refer to actual moment-by-moment signal strength; they are r.m.s. or other values depending on the type of indicator used, in which the whole cycle is lumped together without distinguishing between positive and negative halves. To get the " shape " of the non-linearity—which often affects positive and negative half-cycles differently—it is necessary to plot instantaneous values. This can be done by using positive and negative peak-reading voltmeters, but much more quickly and easily by connecting the input signal to the X plates of a cathode-ray tube and the output to the Y plates—the second method of connection described in Sec. 11.6. To make the X and Y deflections about equal, the potential divider shown in Fig. 11.5 will probably have to be used.

This kind of input/output characteristic is sometimes distinguished from the previous one by the name *transfer characteristic*. If it is a perfectly straight line, as in Fig. 11.9*a*, the unit under test is perfectly linear and the phase difference between input and output is either 0° or 180°—which one it is can be ascertained by taking the Y-plate connection away from the amplifier output and attaching it to the X plate; this shows the 0° direction of slope. Of course 0° and 180° can always be interchanged by reversing one pair of connections. In general, amplifiers, transformers, etc., have zero phase shift at some middle frequency, and the line opens out into an ellipse at low and high frequencies. Shunt inductance and series capacitance cause a phase lead at low frequencies; shunt capacitance and series inductance a phase lag at high frequencies. The actual phase angle can be determined from the ellipse as in Sec. 9.10.

Phase shift makes no audible difference to continuous tones, but is important in television and in negative-feedback amplifiers, and if excessive it has some audible effect on transients. Even when not important in itself it is a symptom of frequency distortion. Phase shift can more easily be detected on the c.r.t. than the accompanying reduction in gain.

When there is no phase difference between X and Y signals, the transfer trace is a line enclosing no area and shows up any non-linearity very clearly—more so than displaying the waveform of the output on a linear time base. Also it is easier to carry out, because no time base is needed, nor is it so important for the input waveform to be perfectly sinusoidal. Fig. 11.9*b* and *c* are examples of how non-linearity appears. But when phase shift makes the trace an ellipse it is not so easy to detect a small amount of non-linearity. In Fig. 11.9*d* the non-linearity combined with phase shift is unmistakable, but a small amount of distortion might go unnoticed, so for this test it is advisable to close the ellipse up by introducing a suitable amount of phase shift into the X deflection. The best system for doing this depends on the circumstances, but the network and values shown in Fig. 11.10 are likely to cover most requirements.

Fig. 11.9*e* shows up another amplifier defect—parasitic oscillation occurring at one part only of the signal cycle. This would cause an unpleasant buzz or blast in sound reproduction, but as the oscillation would not occur when the amplifier was at rest the trouble would be very difficult to diagnose and locate without a c.r.t. This type of test, calling for very little apparatus in addition to an a.f. oscillator and the tube itself, gives more information on an amplifier in a few minutes than hours of testing with meters.

Non-linearity can be examined under the microscope, as it were, by deducting the fundamental from the output, leaving only the distortion products to be applied to the Y plates. These may be amplified until even a very small percentage can be seen clearly and visually analysed. An elaborate equipment for doing this, enabling much information

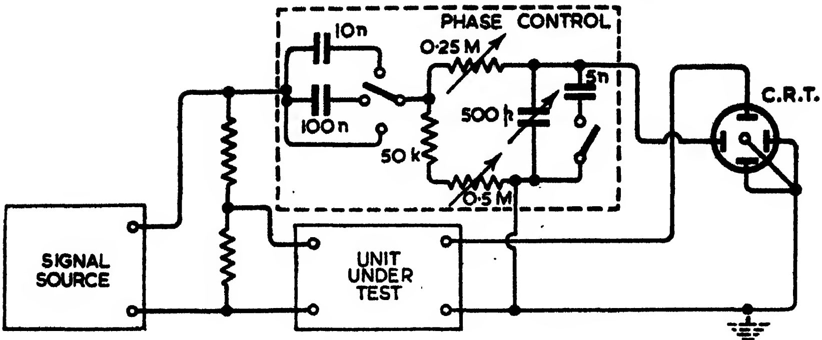


Fig. 11.10—Suggested phase-shift network for facilitating non-linearity measurements by c.r.t.

to be extracted, is described by E. R. Wigan (*W.W.*, June 1953); and a simple arrangement consisting mainly of a parallel-T filter made up from ordinary components, by V. J. Tyler (*W.W.*, Sept. 1953). D. C. Pressey (*W.W.*, Feb. 1954, corrected in March 1954, p. 128), shows still another system, based on computor technique, in which the distorted output is added to the input in opposite phase and equal amplitude. All these methods are (or can be made) more or less quantitative, so might perhaps be included in one of the later Sections, on measurement. But before going on to them one will do well to consider what exactly should be measured.

11.13. BASIS OF NON-LINEARITY MEASUREMENTS

If the purpose of the amplifier is sound reproduction, the result of non-linearity is audible unpleasantness. But one cannot measure unpleasantness as such. The thing to do is to find some measurable feature about non-linearity that is proportional to the resulting unpleasantness. The link-up between objective measurement and subjective hearing can never be precise, because people have different ideas about how unpleasant a thing sounds, and their impressions vary with the kinds of sound being reproduced (Sec. 3.3). However, if a basis for measurement is worked out that lines up with most people's impressions most of the time, then it is reasonable to expect that an amplifier which on that basis—which *can* be precise—shows less non-linearity than another will on the whole be the better amplifier when judged by ear (other things, of course, being equal). If such expectations are often disappointed it shows that the basis for measurement is wrong. But for ordinary purposes it must be practical—not something that necessitates equipment too elaborate for ordinary use, or a voluminous expression of results.

A commonly employed basis is the percentage of harmonics introduced by the non-linearity when a pure single-frequency signal is applied. It is necessary, of course, to specify the frequency and amplitude of the signal. For ease of comparison one would like to be able to express the amount of distortion as a single number. It is possible to specify the total of all the harmonics in the output as a percentage of the fundamental, but that fails altogether to line up with the audible results. For example, 1% ninth harmonic sounds worse than 10% second harmonic. So various attempts have been made to bring them into line by "weighting" the harmonics according to their relative unpleasantness. In one of these attempts, made by the British Radio Manufacturers' Association (as it then was),* the amplitude of the nth harmonic is multiplied by $n/2$ before squaring it, adding the result to those for the other harmonics, taking the square root as the total, and dividing by the unweighted r.m.s. total output (harmonics plus fundamental) to give the so-called distortion factor. According to experiments carried out by D. E. L. Shorter of the B.B.C.,† it appears that a heavier weighting factor—$n^2/4$—gives better agreement with listening tests, but the resulting figure depends on unmeasurably small high harmonics.

The single-frequency test signal does not adequately represent any type of programme that would hold the public interest for long; in real programmes there are many frequencies simultaneously, and non-linearity not only adds harmonics of them all but also whole ranges of intermodulation tones. The lower harmonics at least harmonize, but the corresponding modulation products are generally discordant. It can quite easily be demonstrated‡ that the unpleasantness of non-linearity distortion in typical conditions is due almost entirely to intermodulation, not to harmonics. Except that two signals are required at once, measuring total intermodulation is not very much more complicated than measuring total harmonic distortion. But the fact that most of the unpleasantness is due to intermodulation does not necessarily mean that the figure for total intermodulation is a more reliable guide than the already discredited figure for total harmonics, or that because harmonics contribute little to the unpleasantness they cannot be used as a measure of it,§ for indeed harmonics can be regarded as intermodulation products between one frequency and itself. A great deal has been published on this subject, but it is difficult to arrive at generally accepted conclusions because so much depends on the conditions assumed. There is a need for more research.‖ As a rough sort of summary it is perhaps fair to say that total harmonic distortion (unweighted) is a very unreliable guide to audible distortion and is useful only for comparing amounts of non-linearity of the same type.

* *Proc. Wireless Section I.E.E.*, Sept. 1937.

† "The Influence of High-order Products in Non-linear Distortion." *E.E*, April 1950.

‡ *W.W.*, April 1955, p. 192.

§ "Relations Between Amplitudes of Harmonics and Intermodulation Frequencies", by M. V. Callendar and S. Matthews. *E.E.*, June 1951, corrected in the issue of July 1951, p. 277.

‖ Such as that carried out by E. R. Wigan and recorded in *E.T.*, Apr. and May 1961 ("New Distortion Criterion").

Total intermodulation is a more sensitive test of non-linearity, but there is doubt whether it is much better as a measure of audible distortion, except where the harmonics are outside the frequency band of the unit under test. To investigate non-linearity properly it is necessary to measure separately the different harmonics and the different intermodulation products. A strong 100-c/s signal is liable to produce harmonics at 200, 300, 400, etc., c/s, and if accompanied by another signal at frequency f there will be intermodulation tones at $f \pm 100$, 200, 300, etc., c/s, and perhaps $2f \pm 100$, 200, etc., c/s. Their relative strengths are an indication of the type of non-linearity. For example, if the second harmonic predominates, that indicates a gradual curvature all the same way, as in Fig. 11.9*c*, typical of a single triode; if there is a strong third harmonic there must be a double curvature as at *b*, typical of a single pentode. Even if the distortion is to be expressed finally as a single number, it is necessary to find the values of the separate distortion tones in order to weight them so that the figure gives at least a fair idea of relative unpleasantness.

The subject is discussed more fully by the author (*W.W.*, July 1955), with 26 references to further literature, and a suggested form of distortion test to give results indicative of unpleasantness yet with reasonable brevity. The main idea is to adopt a fixed frequency of intermodulation product somewhere in the most audible part of the frequency band : 1–2 kc/s. Several fixed frequencies could be used for the stronger of the two input tones, at representative points in the a.f. band; say 65, 800, 3,000 and 12,000 c/s. At each of these frequencies a second tone of quarter the amplitude is then swept over the band and the amplitude of the fixed product tone measured at every occurrence. An inexpensive wave analyser, using nearly zero i.f. instead of the conventional supersonic frequency, and a twin a.f. signal generator, were described in the following issue (Aug. 1955).

11.14. CALCULATION OF HARMONICS

If an accurate transfer characteristic is available (Sec. 11.12) the corresponding harmonics can be calculated. It is necessary for the characteristic to be a line enclosing no area; consequently there is a temptation to obtain it at a frequency giving no phase shift through the amplifier—usually a middle frequency. But harmonic distortion is much more likely to be severe at low frequencies such as 50 c/s, and the resulting distortion comes at frequencies where it is most likely to be troublesome. So the phase-shift should be balanced out by means such as Fig. 11.10. Owing to different shifts at the harmonic frequencies, success cannot be guaranteed.

The following method is adapted from that described by J. A. Hutcheson (*Electronics*, Jan. 1936) who demonstrated it for all harmonics up to and including the seventh; and it involves only simple arithmetic. Fig. 11.11 shows how the requisite data are derived from the cathode-ray trace. The total output swing, *a*, is measured, which can most conveniently be done by cutting off the input (horizontal) deflection. The other quantities, b to g, are the distances measured vertically (in the direction of the output deflection) from a straight line drawn between the ends of the curve to selected points on the curve. The relative horizontal distribution of these points is shown. Distances measured downwards from the straight line are called negative; those upwards, positive. These signs must be carefully observed throughout the calculation, but the signs of the final answers are of no great significance. The distances can either be measured in volts, if the screen is calibrated,

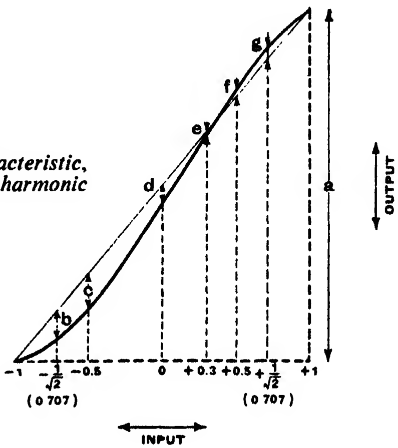


Fig. 11.11—Typical transfer characteristic, showing dimensions needed for harmonic analysis as described

in which case the amplitudes of the harmonics are given in volts; or in any arbitrary units such as millimetres, for the units make no difference to the harmonic percentages.

If the amplitudes of the various harmonics are denoted by V_2, V_3, etc., they are given by

$$V_2 = \frac{f+c}{3} + \frac{d-b-g}{4}$$

$$V_3 = \frac{f-c}{3}$$

$$V_4 = \frac{d-b-g}{4}$$

$$V_5 = \frac{f-c}{3} + \frac{b-g}{2{\cdot}828}$$

$$V_6 = \frac{d}{2} - V_2$$

$$V_7 = \frac{(e - 1{\cdot}82V_2 - 1{\cdot}092V_3 + 0{\cdot}655V_4 - 0{\cdot}699V_5 - 0{\cdot}751V_6).}{1{\cdot}146}$$

The seventh is the only one that takes more than a moment or two to work out, and it can be omitted unless there is reason to believe that the upper odd harmonics are appreciable.

The amplitude of the fundamental, V_1, differs from the amplitude of deflection only by the odd harmonics:

$$V_1 = \frac{a}{2} + V_3 - V_5 + V_7$$

The percentage of any harmonic is $\frac{100\ V_n}{V_1}$ and the percentage of the total harmonics is $\frac{100}{V_1}\sqrt{V_2^2 + V_3^2 + V_4^2 \ldots}$

If weighted by half the harmonic number it is

$$\frac{50}{V_1}\sqrt{(2V_2)^2 + (3V_3)^2 + (4V_4)^2 \ldots}$$

(In an alternative method of reckoning, the total r.m.s. output is substituted for V_1 in these two expressions.)

As an example, suppose that in Fig. 11.11

$$a = 120$$

$$b = -10 \qquad e = -2$$
$$c = -12 \qquad f = 2$$
$$d = -7 \qquad g = 3$$

Then $$V_2 = \frac{2-12}{3} + \frac{-7+10-3}{4} = -3{\cdot}33$$

$$V_3 = \frac{14}{3} = 4{\cdot}67$$

$$V_4 = 0$$

$$V_5 = 4{\cdot}67 + \frac{-10-3}{2{\cdot}828} = 4{\cdot}67 - 4{\cdot}60 = 0{\cdot}07$$

$$V_6 = -\frac{7}{2} + 3{\cdot}33 = -0{\cdot}17$$

$$V_7 = \frac{-2+(1{\cdot}82\times3{\cdot}33)-(1{\cdot}092\times4{\cdot}67)-(0{\cdot}699\times0{\cdot}07)+(0{\cdot}751\times0{\cdot}17)}{1{\cdot}146}$$

$$= \frac{-2+6{\cdot}07-5{\cdot}10-0{\cdot}05+0{\cdot}13}{1{\cdot}146} = -0{\cdot}83$$

$$V_1 = 60 + 4{\cdot}67 - 0{\cdot}07 - 0{\cdot}83 = 63{\cdot}77$$

percentage	2nd	harmonic	=	5·2
,,	3rd	,,	=	7·3
,,	4th	,,	=	0
,,	5th	,,	=	0·1
,,	6th	,,	=	0·3
,,	7th	,,	=	1·3
,,	total	,,	=	9·1
weighted ,,	,,	,,	=	20·2

It is essential to measure the trace accurately and to avoid distortion of it, for unless the electrical distortion is very bad the harmonics are relatively small. Also, because they are calculated as differences between probably larger quantities, the working must be done to at least one more significant figure than is expected in the answers. Generally two significant figures are satisfactory in the answers, and only the first is likely to be at all reliable even if the work has been done very carefully. For this reason c.r.t. traces have to be focused very sharply to be worth using. The same analysis can however be applied to calculated or plotted dynamic valve curves; i.e., valve transfer characteristics inclusive of load impedance. And more accurate results might be obtained by combining this type of calculation with the fundamental-removing technique described at the end of Sec. 11.12.

Calculation of harmonics from waveforms is explained in some of the larger books on a.c., and by R. C. de Holzer (*E.E.*, June and July 1945) and D. R. Turner (*E.E.*, Jan. 1953).

11.15. MEASUREMENT OF HARMONICS

Total harmonic distortion is usually measured by comparing the signal output voltage with and without the removal of the fundamental by a sharp filter or otherwise. In the Marconi TF 142F Distortion Factor Meter, a bridged-T is used for suppressing the fundamental and is then replaced by an attenuator, the setting of which when the meter reading is equalized shows the percentage distortion directly. In this instrument the fundamental can be varied over the range 100 c/s to 8 kc/s. Obviously a great simplification can be made by working at a fixed fundamental frequency, commonly 400 or 1,000 c/s, but this greatly restricts the utility of the instrument, because it is the increase in distortion at the extremes of frequency that is usually of most significance. Details of a simple design of this type, with an added facility for measuring third harmonic only, are given by T. D. Conway (*W.W.*, March 1954).

Although measurements confined to total distortion are easy to make and are expressed by a single number, their correlation with subjective impressions is poor unless something is known or can safely be assumed of how the total is made up. It is therefore a good idea to combine this kind of measurement with the visual observation described at the end of Sec. 11.12.

The ideal in harmonic distortion measurement is to be able to find the amplitudes of all significant harmonics—bearing in mind that small fractions of 1% are significant where high harmonics are concerned. The proper equipment for this is a signal source that does not itself contribute appreciable harmonics—or from which harmonics are tuned out by a filter—and a wave analyser (Sec. 5.23). The test should be made at various amplitudes and frequencies, especially at the lowest frequencies the equipment under test is supposed to handle.

A good wave analyser of the highly selective valve-voltmeter type is a most versatile instrument, enabling intermodulation products, hum, and even random noise to be measured. The snag is that it is expensive and not readily constructed by an amateur. An exception is the " zero-beat " type developed by the author (*W.W.*, Aug. 1955).

Though less convenient in use, a method of wave analysis described by D. Martineau Tombs (*W.E.*, July 1950), is inexpensive, for it can be carried out by means of apparatus that is available in most laboratories or can easily be made up. The scheme is to use a negative resistor such as a dynatron to adjust the Q of a resonant circuit tuned to each harmonic in turn. The voltage of any harmonic is inversely proportional to the Q needed to bring it up to a uniform level, so it is directly proportional to ΔC in the capacitance-variation method of Q measurement (Sec. 9.28). The formula is

$$e_n = \Delta C \, \omega_n^2 \, LE/2$$

where e_n is the voltage of the nth harmonic, ω_n is 2π times its frequency, L the tuning inductance, and E the standard level, observed on an oscilloscope. It is a particularly interesting experimental method.

A survey of commercial wave analysers and distortion-factor meters has been made by R. Brown (*B.C. & E.*, Oct. 1960).

11.16. MEASUREMENT OF INTERMODULATION

Here again there are methods for measuring total distortion and others that distinguish between the various intermodulation products. The latter are much to be preferred, for experience shows that equal amounts of total intermodulation differ widely in unpleasantness according to how the total is made up (J. H. O. Harries, *W.E.*, Feb. 1937). For instance, if an amplifier carries a strong 50-c/s signal and a weaker one at 1,000 c/s, a given amount of second-order intermodulation (950 and 1,050 c/s)* is less objectionable than the same amount of third-order (900 and 1,100 c/s). These generally correspond to prominent second and third harmonics respectively.

Because of the insensitiveness of the ear (and perhaps the loudspeaker) to low frequencies, a very low frequency, while not itself conspicuous, may be strong enough at some stage to cause serious mutilation of higher frequencies; this is one of the commonest causes of bad reproduction. Moreover, non-linearity is likely to occur at a smaller amplitude at low frequencies than at medium. So it is most often recommended that the two signals applied shall consist of a strong low frequency and a weaker medium or high frequency. But there is an alternative scheme (explained by A. P. G. Peterson in *G.R.E.*, March 1951), in which variable frequencies of equal amplitude and differing by a fixed frequency such as 1,100 c/s are used. For general purposes, however, the former is more revealing and less likely to cause difficulty.

Some care is needed in bringing the two signals to the input, to ensure that the signal sources do not intermodulate one another. A convenient source of the low frequency is the ever-useful 50-c/s mains, and

* This is sometimes called first-order intermodulation.

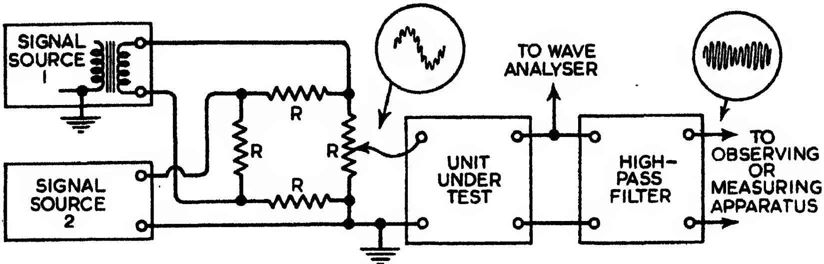


Fig. 11.12—Method of connecting two a.f. signal generators for intermodulation tests. The R bridge is used to prevent the generators from affecting one another

this is hardly likely to be seriously affected by a weak 3,000-c/s (say) signal. If the source of the latter has a large amount of attenuation in its output, it too may be safe even if the two signals are simply applied in series. To be quite sure, the bridge method shown in Fig. 11.12 can be used. At least one of the signal sources should be connected through a transformer to avoid earthing difficulties. It is usual to arrange that the high-frequency signal is 12 dB below the low (i.e., one-quarter of its voltage); the combined input level can then be adjusted as shown. The amplifier should of course be working into its usual load.

If a wave analyser is available, the rest is easy: this instrument is itself sufficiently selective to measure the various output components separately in turn. The amplitudes of the modulation products—second-order, 2,950 and 3,050; third-order, 2,900 and 3,100; etc., in this example—are expressed as percentages of the 3,000-c/s signal output. Components at 6,050 c/s, etc., should also be looked for. See Sec. 11.13 for references to the author's inexpensive wave analyser and recommended intermodulation test scheme.

When a wave analyser is not available it is necessary to use a high-pass filter (Sec. 14.26) to remove the low-frequency signal and its harmonics. In the present case a cut-off at about 1,000 c/s would be suitable. The greater the ratio between the two input frequencies, obviously the easier it is to effect a clean separation without endangering any of the significant modulation frequencies.

What is done next depends on how the distortion is to be observed. Removal of the strong signal enables the damage to the weak one to be examined on an enlarged scale by amplification, as with harmonics in Sec. 11.12. Fig. 11.13 shows the simple requirements for visual and/or aural observation. If the amplifier under test is linear within the peak-to-peak limits of the strong signal, the output appears at uniform amplitude, as at Fig. 11.14*a*. If there is non-linearity, its nature is revealed by the envelope of the high-frequency signal. Typical single-triode or second-harmonic or second-order modulation distortion is shown at *b*; the " carrier wave " goes through one cycle of amplitude variation per cycle of the low frequency. Trace *c* shows similar

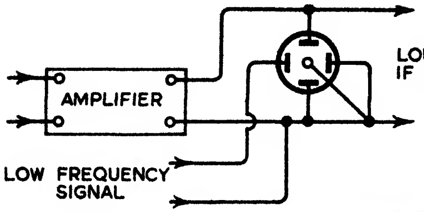


Fig. 11.13—Connection of c.r.t. for observing intermodulation. Typical traces are shown in Fig. 11.14

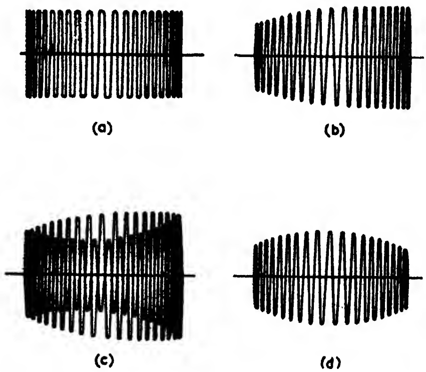


Fig. 11.14—Intermodulation traces showing (a) *linearity (no intermodulation),* (b) *second-order intermodulation,* (c) *same with phase shift, and* (d) *third-order intermodulation*

distortion with phase shift. At *d*, characteristic of pentode distortion, there are two cycles of modulation per low-frequency cycle, indicating a pronounced third-order modulation. The distortion can be expressed numerically as the percentage depth of modulation (Sec. 11.23) for a stated low-frequency input.

Alternatively, the output from the high-pass filter—amplified if necessary—may be applied to a diode rectifier, the high frequency removed by a low-pass filter, and the amplitude of the low-frequency output, which represents the distortion, measured by a valve voltmeter. Either the same meter switched, or a separate one, is used to measure the input to the rectifier, and the ratio of one to the other, multiplied by 100, is the percentage modulation. To give a true reading of the total, the meter ought to read r.m.s. values. Some care is needed with this method to avoid errors due to the rectifier, and for accurate results allowance must be made for its efficiency being less than 100%. Fuller information is given by E. W. Berth-Jones (*W.W.*, June 1951), and design data for a simple intermodulation meter are given by J. M. van Beuren (*Audio Engineering*, Nov. 1950). This total intermodulation figure is obviously less informative than the separate values obtainable with a wave analyser.

11.17. A.F. MODULATION HUM

In certain circumstances, such as push-pull operation, the modulating tone may be the ripple left over as a result of inadequate smoothing. In a straight amplifier this ripple would show up as excessive hum, but in push-pull the hum is cancelled out. If too much advantage is taken of this to reduce smoothing components, the ripple, though sufficiently balanced out to be inoffensive with no signal, may nevertheless swing each valve over so much of its characteristic as to modulate a signal. This modulation hum can be identified by examining the output of the amplifier with the cathode-ray tube on a base of the a.c. supply feeding the amplifier power unit. A single test signal, of a relatively high frequency, is applied; and a filter is not essential, so the apparatus is very simple. Modulation hum shows up as an irregularity in the envelope of the signal wave, and its seriousness can be gauged by comparison with the mean amplitude of the signal (Sec. 11.23).

11.18. MEASUREMENT OF HUM

The measurement of straight hum might seem quite simple, but it is not. The results of using an output power meter or an a.c. voltmeter across the output of the amplifier are almost entirely useless. Owing to the fact that the sensitivity of the ear depends enormously on frequency, and owing also to such complications as speaker resonances, a barely audible low-pitched hum may give a pronounced reading on a meter which fails to show any response at all to a high-pitched ripple only too obvious to the listener.

One solution is to interpose a weighted network—a filter designed to attentuate frequencies in proportion to the sensitivity of the ear—followed by or combined with an amplifier. Details of a hum voltmeter are given in Sec. 5.23 (3). When connected across the load, R_L, of an amplifier under test, the voltmeter reading divided by $\sqrt{R_L}$ gives a comparative indication of the sound that would be produced by a perfect loudspeaker substituted for R_L. Note that the weighting amplifier gives a nearly level aural characteristic over the whole audible frequency band, and any hiss or other noise present is also indicated; if only hum is to be included, a filter cutting off above about 1,000 c/s should be interposed. It should be noted, however, that if the hum is being introduced via stray capitance, an appreciable proportion of the audible result may be due to harmonics at frequencies above even 1,000 c/s.

11.19. SQUARE-WAVE TESTS

In the type of amplifier test described in Sec. 11.8, resulting in a response graph on a frequency base, each point is observed after a signal having a single frequency has been applied for a practically infinite time (say a second or two!) so that the response has reached its steady-state value. There is an alternative, in which the signal comprises a very wide range of frequency within a very short space of time, and the response is shown on a time base. This is clearly a closer approach to most real types of audio programme than the other. The signal shown in Fig. 11.15 (called the Heaviside unit function, or step voltage) includes all frequencies, if the transition from one constant level to another takes place instantaneously, but of course this is only a theoretical form. A practical approximation is a periodic square wave with a fundamental frequency near the lower end of the amplifier band, and a sufficiently steep wave front to include frequencies up to the top end of the band. For square-wave generators see Sec. 4.38.

The output of the amplifier or other unit under test is observed by an oscilloscope with its linear time base synchronized to about one-third of the square-wave frequency so that at least one unbroken cycle of the output can be seen. A perfect amplifier would of course show a replica of the input waveform, which should be as nearly as possible

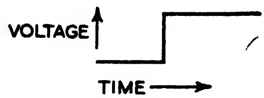


Fig. 11.15 (above)—Step voltage, which is the basis of transient tests

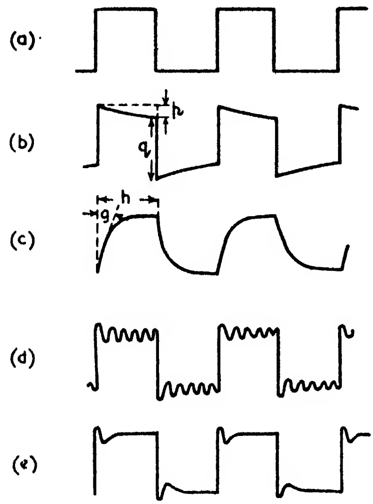


Fig. 11.16 (left)—Characteristic traces obtained in square-wave tests, from which the performance of the amplifier under test can be inferred

a perfect square wave (Fig. 11.16*a*). Drooping tops, as at *b*, are caused by the phase advance that is the usual sign of a falling-off in response at the low frequency.

Suppose for example that it is caused by one capacitance coupling of the usual type, in which a constant voltage input is applied to C and R in series and the output taken from across R. Assuming that the droop, measured by p/q, is not more than about 0·3, and denoting the square-wave frequency by f_s, the cut-off frequency (at which the coupling attenuates to the extent of 3 dB) by f_c, and the phase shift at f_s by ϕ:

$$\tan\phi = \frac{f_c}{f_s} \simeq \frac{2p}{\pi q} = 0{\cdot}64\frac{p}{q}$$

If the c.r.o. trace is narrowed by reducing the time-base frequency, a p/q ratio as small as 1/30 can quite easily be observed, so that a cut-off frequency of, say, 30 c/s can be found with f_s as high as 700 c/s. It has been pointed out that strictly one cannot infer anything about the performance of an amplifier at any frequency lower than that of the square wave, and of course that is perfectly true if there is no knowledge of the circuit causing the low-frequency fall-off. If there were several capacitive couplings and perhaps inductive shunts the law would be different. But the relationship $\sin \phi = 1{\cdot}27\ p/q$ holds good so long as the sloping top is straight, and is useful for measuring phase shifts up to about 12°, which are accompanied by attenuation less than 0·1 dB.

The phase lag due to falling-off at the high frequencies shows up as at *c*. Assuming the cause is one shunt capacitance (or series inductance), and that the half-cycle of f_s lasts long enough for the output to reach practically the full amplitude, the − 3 dB frequency, f_c, can be found by measuring the *initial* rate of rise, as shown:

$$\tan \phi = \frac{f_s}{f_c} \simeq \frac{\pi g}{h}$$

For g to be measurable, without conflicting with the assumption, f_s for this test ought to be about 0·3–1 times f_c—say 5 kc/s for typical a.f. amplifiers. In both these tests it is necessary for the positive and negative half-cycles to be equal.

A good idea of the bandwidth and phase shift of a system can therefore be obtained by inspecting the output waveforms at two frequencies. But these waveforms show a good deal more, notably certain types of transient distortion. A picture such as *d* indicates a pronounced resonance and its frequency—in this example $10f_s$, for there are ten cycles of lightly damped oscillation per square-wave cycle. To detect such a resonance with a frequency test it would be necessary to sweep over the whole frequency band, looking for a sharp peak; and even when found it would not give such a good idea of the distortion. If the circuit responsible is so heavily damped as to be almost non-oscillatory, the appearance is as at *e*.

There are very many other possible shapes, and their interpretation is a fascinating and instructive study, which can be pursued by trying different types of circuit, preferably with variable components.

If negative-feedback amplifiers are subjected to square-wave tests, care must be taken that the wave-front does not rise faster than the amplifier can respond. If it does, the amplifier may be overloaded and the feedback put out of action at a signal amplitude that could easily be handled at the fundamental frequency. This danger has been pointed out and explained by W. T. Cocking (*W.W.*, March 1946) with particular reference to the cathode follower (Fig. 11.17). (See also "Cathode Follower Distortion", *W.W.*, May 1961.) It is generally supposed that one of the advantages of the cathode follower is its very low output impedance, by virtue of which it can work into a relatively large capacitance without serious attenuation or distortion. And this is true so long as the input voltage does not change too fast. If it does, the capacitance *C*—even if it is only unintentional stray—prevents the cathode potential from following that of the grid quickly enough, and if the signal is negative-going the valve may be momentarily cut off, with resulting distortion.

Amplifiers that are to be used for steep-fronted waveforms, as in television, should be tested with such waveforms or their equivalents, but it would be unfair to use much steeper test signals. Still more so would it be to test a.f. amplifiers with such signals; square waves, if nearly perfect, should be smoothed off somewhat; say by a top-cut

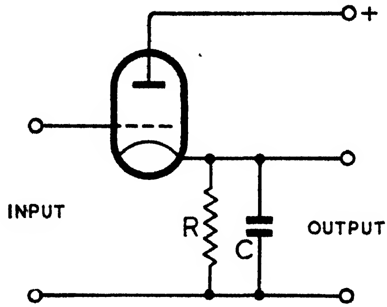


Fig. 11.17—Cathode follower, showing capacitance C *which is always present whether intentional or not and which restricts the frequency or amplitude of the applied signal*

CR Circuit in which $1/CR$ is about 30 times the highest frequency for which the amplifier is required.

11.20. GRAMOPHONE PICKUPS

Most of the methods already described for testing amplifiers can be adapted for pickups by using special test records, listed in Table 14.23. These provide single-frequency signals, both as constant spot frequencies and continuously varying, for frequency characteristics, and two-frequency signals for intermodulation. Triangular-wave records, which ideally would produce a square-wave output, have been produced experimentally but at the time of writing do not appear to be on sale, presumably because of excessive wear.

For most tests the output of the pickup is unlikely to be sufficient, and if the characteristics of the pickup itself are required it is necessary to use an amplifier without appreciable distortion of the kind being investigated. The alternative is to test the pickup with its own amplifier; the characteristics of the amplifier alone can be measured and those of the pickup deduced from the two sets of results.

Intermodulation tests are described by S. Kelly (*W.W.*, July 1951) and discussed by L. J. Elliott (*W.W.*, Sept. 1951).

11.21. V.F. AMPLIFIERS

The testing of v.f. amplifiers is an extension of a.f. amplifier-testing technique, to cope with frequencies up to several Mc/s. Because of the place of pulse signals in television, and the visual nature of the final presentation, there is a greater emphasis on square-wave methods.

It is possible, of course, to take conventional frequency characteristics, and there are special oscillators designed to cover the whole v.f. band (Sec. 4.27). Care is needed to ensure that there is no undesired loss of signal at the high-frequency end. Everything is done at as low an impedance as possible, to minimize the effects of unavoidable shunt capacitance. The cathode follower is a much used device for obtaining a low-impedance output from a high-impedance source. If it is necessary to use a screened input lead, the extra capacitance introduced thereby can be very greatly reduced by using the double-sheath system due to A. D. Blumlein shown in Fig. 11.18 and described by V. H. Attree (*E.E.*, March 1949). But the added capacitance across the

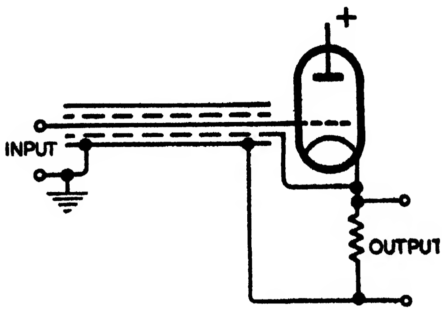


Fig. 11.18—Cathode-follower device for reducing the input capacitance of a screened lead

output accentuates the risk of overloading referred to in Sec. 11.19, and it is necessary to use a low load resistance and not too steep a wavefront.

Square-wave generators for v.f. must be capable of giving a good waveform at both ends of the frequency scale; this necessitates full use of the usual wide-band techniques. For equipment, procedure and interpretation of results, see T. B. Tomlinson (*E.E.*, June 1949); for more detailed analysis of the v.f. amplifier by this method, see P. M. Seal (*Proc. I.R.E.*, Jan. 1949); and for general television receiver testing, see M. V. Callendar (*W.W.*, Feb. 1952).

For testing television v.f. lines and equipment the standard instrument is a " pulse and bar " generator (Pye PTC 1201) which yields the waveform shown in Fig. 11.19. The imperfections of the channel are judged by the extent to which the waveform seen at the far end on an oscilloscope differs from the known input form from the generator. The bar and sine-squared pulse show up low-frequency and high-frequency imperfections respectively, as well as ringing and echoes. The pulse can be switched to a width (at half-peak amplitude) of either 0·17 microsecond, which provides a spectrum well maintained up to 3

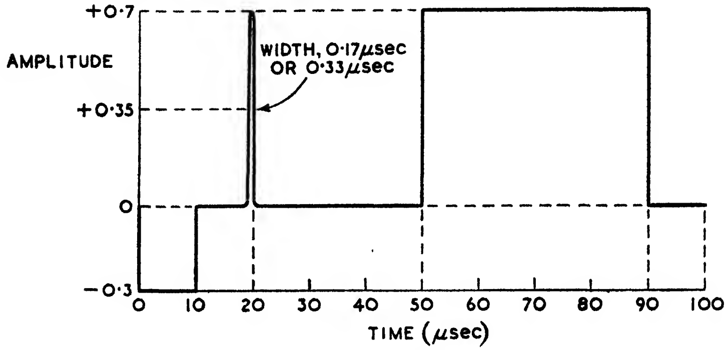


Fig. 11.19—Standard pulse-and-bar waveform used for testing television channels and equipment

Mc/s and cutting off rapidly thereafter, or of 0·33 microsecond. There is a standard code for expressing the visual distortion in K numbers, measured with a special graticule on the oscilloscope screen. For instance, reduction in amplitude of the pulse as compared with the bar indicates deficiency of top-frequency response at 3 Mc/s or 1·5 Mc/s, according to the pulse-width used. The subject is treated fully, with numerous oscillograms of actual responses, by N. W. Lewis (*Proc. I.E.E.*, Part III, July 1954). A brief survey of waveform testing methods for television links is given by A. R. A. Rendall (*E. & R.E.*, Dec. 1957).

11.22. R.F. AMPLIFIERS: GAIN

In this section and the next, " r.f." should be understood to include " i.f." When measuring r.f. amplifiers it is inadvisable—unless the frequency and gain are quite low—to switch a meter from input to output, or to bring the signal at the high-level end of the attenuator out into the open. The attenuator is almost invariably ahead of the amplifier and incorporated in the source, which thus becomes a standard-signal generator. Its output impedance is made as small as possible—usually about 10 Ω—so when it is connected in parallel with a coupling tuned to the signal frequency it effectively short-circuits the coupling, and the signal voltage is not appreciably affected. It should be remembered that whereas for valve amplifiers a constant-voltage signal source is right, transistors normally require a constant-current source. The usual type of signal generator is converted thereto by inserting series resistance large compared with the input impedance of the transistor. Remember too that transistor r.f. amplifiers can easily be overloaded by a signal generator, so the signal level must be considered carefully.

It is seldom that one can assume that the impedance of even a valve voltmeter is high enough, and its stray admittance low enough, not to affect the signal where it is connected. But the amplifier being tested is almost certain to work into either a detector or another amplifier followed by a detector, and if there is not already some sort of signal-level indicator it is usually quite easy to extemporize one; for instance, by putting a microammeter in series with the load resistor of the detector. The rectified current in a diode detector is only a few microamps, but if a low-reading microammeter is not available an alternative is to observe the anode current in an added valve used as a z.f. amplifier. The popular double-diode-triode can usually be slightly modified to respond to the z.f. component of the diode's output. Or, if the signal is modulated to a constant depth, the output level can be measured at a.f. by an ordinary rectifier voltmeter.

Output indicating is simplified, and possible non-linearity between

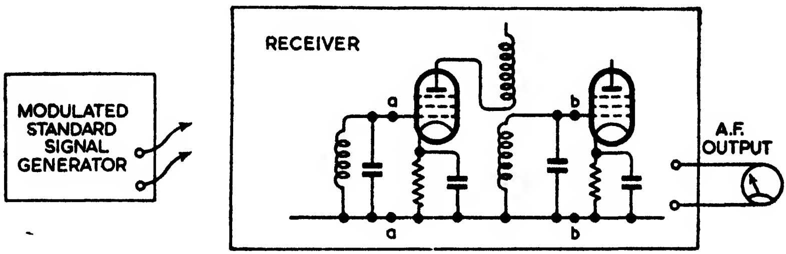


Fig. 11.20—Method of measuring the gain of an r.f. stage, by applying a signal generator in turn to aa *and* bb

the output of the part under test and the indicator is kept out of the results, if the indicator is used merely to show equality, the gain being read on the attenuator. Fig. 11.20 shows a typical case: to measure the gain of an r.f. stage between *aa* and *bb* the signal source is first connected to *aa* and its output adjusted to give a suitable indication. It is then transferred to *bb* and its output increased to give the same indication. The ratio of increase is the required gain, for the level at *bb* must have been the same both times. The gain as measured includes the influence of any feedback from parts of the apparatus following *bb*.

11.23. R.F. NON-LINEARITY DISTORTION

Non-linearity distortion of a carrier wave as such is of less importance than distortion of the modulated envelope. The severity of envelope distortion normally increases with the depth of modulation. For observing it a modulated oscillator is of course needed, and the modulating signal should be accessible for connecting to the X plates of a cathode-ray tube. The depth and linearity of modulation of the oscillator can be seen by connecting the modulated output to the Y plates, if necessary through an amplifier known to be reasonably distortionless. The result is a trace such as Fig. 11.21*a*. The depth of

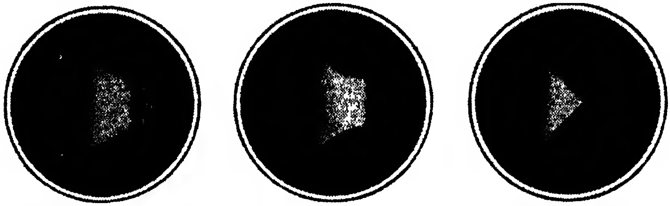


Fig. 11.21—C.r.t. traces from which the purity and depth of modulation can be determined

modulation is 100 $(A - B)/(A + B)$ %, and absence of distortion is indicated by the straightness of the slopes. These are, in fact, modulation transfer curves (Sec. 11.12). An example of distorted modulation is shown at *b*, and over-modulation at *c*. Having ascertained that the oscillator itself is beyond reproach, one can then apply its output to the amplifier under test and examine the r.f. output of the latter similarly. Note that it is not essential for the waveform of the modulation to be purely sinusoidal in this test.

The c.r. trace has to be exceptionally finely drawn to enable the non-linearity to be measured as in Sec. 11.15. If a detector is used to obtain a m.f. output, this can be analysed for non-linearity by the other methods already described. Such tests of course include any distortion due to the detector.

For measurement of r.f. selectivity, etc., see Secs. 11.27 to 11.40 on receiver tests.

11.24. DETECTORS

The following are the characteristics most likely to be of interest:

Loading. The effective load resistance imposed by the detector on the preceding tuned circuit can be measured by the resonance methods for parallel resistance described in Secs. 9.26 to 9.33; but it should be borne in mind that the loading varies with r.f. amplitude, and that normal working conditions should be reproduced accurately for the test. These considerations both apply especially to the double-diode arrangement for detection and a.g.c., in which the loading resistance falls when the r.f. amplitude exceeds the a.g.c. bias.

Efficiency. This is the ratio of the actual modulation-frequency response to the response of a perfect detector. With a diode detector, for example, this ideal output is equal to the r.f. input voltage multiplied by the depth of modulation; 5 V modulated 30% applied to a perfect detector would give $5 \times 0{\cdot}3 = 1{\cdot}5$ V m.f. The depth of modulation can be measured as described in Sec. 11.23; and, provided that the selectivity between the source and the detector is not sufficient to alter the depth by discriminating against the sidebands, there is no objection to measuring it at the generator, where the voltage is conveniently high. The actual m.f. output and r.f. input voltages can be measured by a suitable valve voltmeter. Detector efficiency varies with input voltage and to some extent with carrier and modulation frequencies. There is normally a loss of efficiency at the highest modulation frequencies (owing to *RC* time constant), and at very low carrier frequencies (owing to insufficient capacitance), and at very high carrier frequencies (owing to stray capacitance, etc.). To investigate the falling-off in efficiency as the input is reduced much below 1 V, a valve voltmeter with pre-detector amplification is desirable.

Non-linearity. A static input/output curve, plotted point by point, is not a safe guide to non-linearity, because the reduced load resistance to a.c. is a possible cause of distortion (though less than is sometimes supposed*). Curves of z.f. output against r.f. input voltage (modulated and unmodulated) are, however, instructive as regards a.g.c. Non-linearity can be observed or measured by the same methods as for amplifiers, with varying r.f. amplitude and modulation depth, but of course it is necessary to check the signal generator for absence of distortion, especially at the maximum depth of modulation. In measuring the m.f. output it is necessary to make sure that the r.f. component has been thoroughly filtered away or it will upset the readings. This can easily be checked by shutting off the modulation and seeing whether the reading is zero.

11.25. FREQUENCY CHANGERS

The problem of the frequency changer is very similar to that of the detector, but is complicated by the presence of the heterodyne oscillator, adjustment of which influences the performance. The most practical test is to measure the heterodyne voltage over the whole of its frequency range to make sure that it nowhere departs too drastically from that recommended for the type of valve. The valve voltmeter ought to cover a range of, say, 2 to 50 V. And of course it is essential that connecting it to the appropriate electrode should not affect the voltage seriously. Conversion conductance can be determined in the straightforward way by applying a known signal voltage to the control grid and measuring the i.f. output current under actual working conditions, but the latter part of it presents some difficulties. The i.f. must be separated as well as possible from the components of other frequencies, by choosing an i.f. widely different from the others and measuring the output voltage with a low-loss valve voltmeter across a sharply-tuned resonant circuit, but the result is still not known unless the resonant impedance of this circuit is known.

Although it is open to the objection that it does not represent working conditions, a method described by Benjamin, Cosgrove and Warren (*Proc. Wireless Section I.E.E.*, June 1937) is delightfully simple to carry out, as both signal and oscillator voltages can be derived from the 50-c/s supply and the indicator is a d.c. milliammeter. Fig. 11.22 shows the arrangement applied to a heptode valve. The electrodes are fed with their appropriate voltages, and the oscillator

* " Diode-detector Distortion ", by W. T. Cocking. *W.W.*, May 1951.

grid receives a signal at the normal oscillator voltage—say 10 V—from the low-frequency source through a grid capacitor appropriate to the very low frequency; 0·1 μF or over would do. The control grid receives a relatively small signal, say 0·35 V, from the same source. The z.f. anode current is read first with this signal in one phase and then with the phase reversed by means of the switch shown. The difference in mA between the two readings, divided by twice the peak signal voltage, is the conversion conductance in mA/V. For further information see *Radio Designer's Handbook*, 4th edn. (Iliffe), pp. 109–111.

For a complete stage, from the grid of the frequency-changer valve to the grid of the i.f. amplifier valve following, the most important

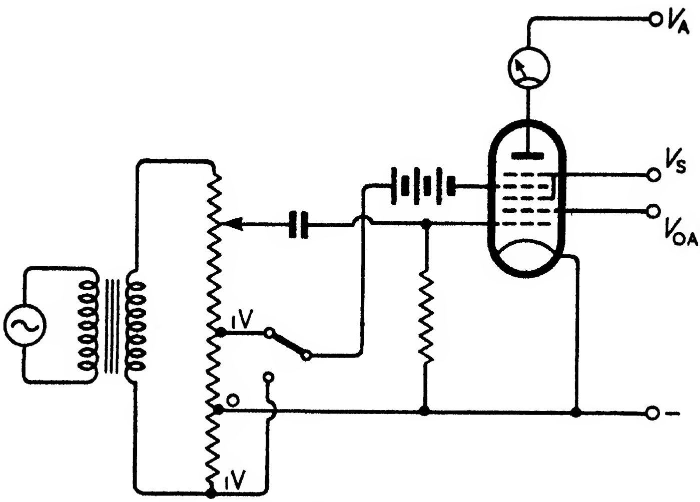


Fig. 11.22—Circuit used for measuring the conversion conductance of a frequency-changer valve, employing one low-frequency source only

characteristic is *conversion gain*. In B.S. 2065 : 1954 this term is restricted to a power ratio, but it is commonly measured as a voltage ratio. This can conveniently be done by the method shown in Fig. 11.20 for r.f. amplifiers, modified by shifting the signal-generator frequency from the chosen r.f. when connected to *aa* to the appropriate i.f. when connected to *bb*. Obviously the depth of modulation must be the same for both readings. To take into account the lowered input resistance of the frequency changer at v.h.f., the measurement should be taken to a point that includes the preceding tuned circuit, and comparison made with the gain at lower r.f.

11.26. POWER UNITS

It will hardly be necessary to remind the reader of the need, when measuring output voltage, to allow for the current taken by the voltmeter, if that is appreciable. This is especially important with e.h.t. supplies for cathode-ray tubes, etc. The subject is dealt with in Sec. 10.1.

If the output voltage is measured at several currents, it is possible to draw a curve connecting output current and voltage, known as a regulation curve. There is no need for many points, because the " curve " is usually indistinguishable from a straight line. Voltages at full, half, and no load are generally enough, except with choke-input smoothing systems, whose characteristic curves take a sharp upward turn near zero load.

When a number of outputs are available in one unit it may be desirable to obtain cross-regulation data, to show the effect of varying load across one output on the voltage supplied to another. If there are numerous outputs one can amuse oneself for quite a long time over this.

11.27. RECEIVERS: SCOPE OF TESTS

The remainder of this chapter, devoted to receiver measurements, in spite of its importance need not extend to great length, because much has already been covered by the preceding sections devoted to the various divisions of receivers.

The nature and extent of tests performed on a receiver naturally depend on whether they constitute a routine check or a detailed analysis; whether, in other words, they are service or production tests, or laboratory measurements. The latter are mainly in view here.

Standard schedules of receiver tests have been drawn up from time to time,* and some conditions and methods have become generally accepted. While for the sake of comparison with results obtained at other times or places it is desirable that one's tests should conform to standards whenever they are appropriate, it is a mistake to bind oneself too rigidly to them. The most important question is: Does the proposed test tell me what I want to know? There is no point in going through a prescribed procedure if it does not. Admittedly, the tests that can be done are limited by the equipment available, so it may not be possible to obtain exactly the required information, but the ultimate purpose of laboratory tests should always be kept well to the fore.

* " Proposed Test Procedure for F.M. Broadcast Receivers ", by D. Maurice, G. F. Newell and J. D. Spencer. *E.E.*, March 1952.

Recommended Methods of Measurement on Receivers for Amplitude-modulation Broadcast Transmissions (Publication 69), *Recommended Methods of Measurement on Receivers for Frequency-modulation Broadcast Transmissions* (Publication 91), *Recommended Methods of Measurement of Radiation from Receivers for A.M., F.M. and Television Broadcast Transmissions* (Publication 106), and *Recommended Methods of Measurement on Receivers for Television Broadcast Transmissions* (Publication 107), Central Office of the International Electrotechnical Commission, Geneva, Switzerland (1954). Obtainable through B.S.I. (see p. x).

See also " The Testing of Communication-type Radio Receivers ", by W. J. Bray and W. R. H. Lowry. *J.I.E.E.*, Part IIIA, 1947. Discussion: *J.I.E.E.*, Part III, 1948.

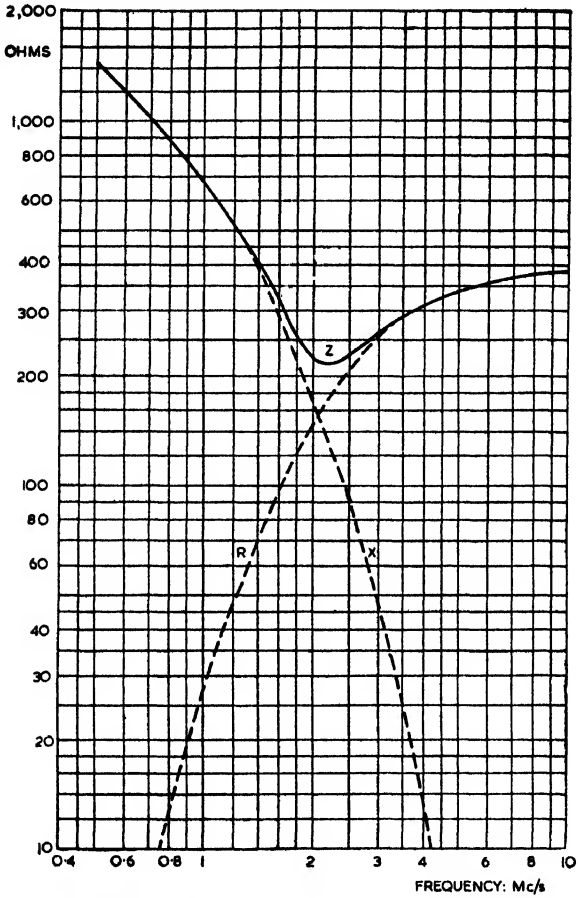


Fig. 11.23—Series resistance, reactance, and magnitude of impedance of standard dummy aerial (Fig. 4.36) as a function of frequency

The output of sound receivers should ideally be measured as sound, but it is very difficult to do this under conditions that give results indicating the true standard of performance. Not only is the equipment very elaborate, but even when it is all available there is no general agreement on the measurement procedure or the interpretation of the results. Reluctantly, then, the loudspeaker is here excluded and the output assumed to be measured in electrical form. When interpreting the measured characteristics it must never be forgotten that they are subject to this omitted factor, which is liable to make the a.f. characteristics either worse or better than they appear, and upset comparisons of sensitivity.

The performance of television receivers can be assessed more reliably, for the cathode-ray tube is not only the normal presentation

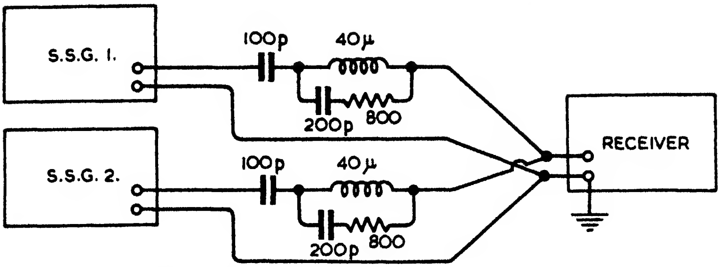


Fig. 11.24—Method of connecting two r.f. signal generators for selectivity measurements. Note that the dummy aerials have twice the desired microvoltage

of the receiver but is also (in an oscilloscope) a reasonably precise measuring instrument.

11.28. STANDARD CONDITIONS FOR RECEIVER TESTS

The following conditions have for the most part become generally accepted and will be assumed unless otherwise stated. Fig. 4.36 shows the specification of the standard dummy aerial connected between signal generator and aerial terminal when testing broadcast sound receivers, and Fig. 11.23 shows its series impedance. It can easily be made up in compact form, and represents tolerably well the characteristics of typical broadcast receiving aerials. For some tests it is necessary to provide signals from two generators simultaneously, and assuming one terminal of each is earthy they should be connected as shown in Fig. 11.24, with double-impedance dummy aerials, *and each generator set to give double the required output.*

Television and other v.h.f. receivers are usually designed to operate from a 75-Ω feeder. The signal generator should then be connected

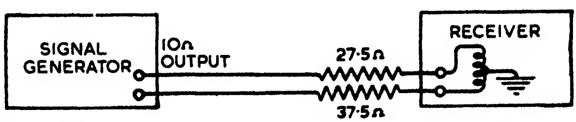


Fig. 11.25—Method of connecting unbalanced signal-generator output to balanced input

through a resistance, sufficient to make up its output resistance to 75 Ω, straight to the receiver r.f. input socket, via a length of 75-Ω coaxial cable if necessary. If the receiver has a balanced input, the source output should also be balanced, as for example in Fig. 11.25.

Injection of the signal into an inductor aerial is possible by connecting the generator directly in series with the aerial; but this has the disadvantage of necessitating a break into the aerial wire, and of putting probably 10 Ω or more into the aerial and so altering its characteristics materially. The alternative is to couple it inductively by means of a coil connected straight to the signal generator (Fig. 11.26). If this coil has N turns of radius R cm and an inductive reactance X ohms, and is situated on the same axis as the receiver aerial and D cm away from it along that axis, then the field strength in microvolts per metre at the receiver aerial when the signal generator output is V microvolts is

$$\frac{18{,}850\, N\, R^2\, V}{[D^2 + (H/2)^2]^{3/2}\, X} \qquad [D \not< \text{double } H,\ W \text{ or } 2R]$$

A suitable coil is 10 cm in diameter and 6 cm long, wound with 20 turns to an inductance of 40 μH and shielded by a wire cage

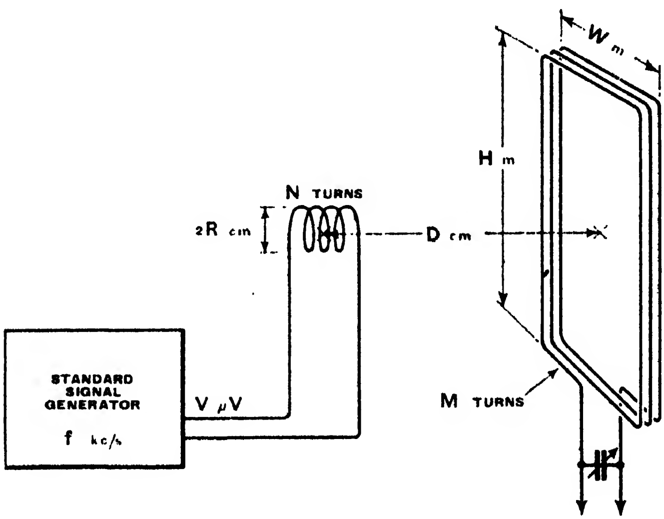


Fig. 11.26—Preferred method of introducing a known signal into a frame aerial. The signal strength is calculated from the quantities shown

arranged so that there are no closed conducting loops in planes at right angles to the axis of the coil. The connecting leads are also to be screened, but the shunt capacitive reactance should be large compared with the inductive reactance, which itself should be so large compared with the impedance of the source that its output should not be seriously lowered by the load imposed by the coil. If the distance, D, between centres of this coil and the receiver aerial is 2 metres, as recommended, the above equation reduces to

$$\text{field strength} = \frac{4{\cdot}67\, V}{f} \text{ microvolts per metre,}$$

where f is the frequency of the signal in kc/s.

If it is desired to know the actual voltage induced in the receiver aerial, the field strength obtained as above must be multiplied by the effective height of the aerial, for which formulae for both frame and ferrite aerials are given in Sec. 14.28.

The voltage developed across the aerial at resonance is, of course, Q times the voltage induced as described.

Standard modulation frequency and depth are 400 or 1,000 c/s and 30% respectively. The author's custom is to make all measurements on sound receivers at a signal frequency of 1 Mc/s unless otherwise required. Suitable frequencies for measuring sensitivity and selectivity in the medium-frequency broadcast band are 600, 1,000 and 1,400 kc/s, supplemented if necessary by 800 and 1,200 kc/s; and in the low-frequency band 175, 225 and 275 kc/s. As for the high-frequency broadcast bands, these are comparatively narrow, and a single frequency in each should be enough; the standard points are 6·1, 7·2, 9·6, 11·8, 15·3, 17·8, 21·6 and 25·8 Mc/s.

A standard output for sound is 50 mW in a resistive load equal to the magnitude of the loudspeaker impedance at 400 c/s or alternatively the load resistance for which the receiver was designed. This output is low compared with the maximum usually available, but an advantage is that it is usually small enough not to bring the a.g.c. into action. For receivers with an output of 1 W or over, 0·5 W is sometimes specified as the standard output, and, for low-power receivers, 5 mW; so it is necessary to mention which has been adopted.

A standard output for vision is 10 V peak at the control electrode of the c.r. tube, with a carrier wave modulated 30% by 100 kc/s.

For measurements the output meter (Sec. 5.7) should be connected in place of the loudspeaker. Whether the output transformer is regarded as part of the receiver or part of the load (loudspeaker) is a matter for discretion—generally the former is preferable—but the distinction should be drawn, for commercial output transformer efficiency is sometimes rather low. If there is no convenient means of disconnecting the speaker, or it is desired to retain it for monitoring, the output readings will be lower than with the meter alone. Assuming that the meter impedance is equal to that of the speaker, an output of 50 mW gives a reading of about 22 mW if the output valve is a triode and 15 mW if it is a tetrode or pentode; but if negative feedback is in use the reduction is less. Both loads should not be used together when taking a.f. characteristics, for the loudspeaker impedance varies with frequency and the load is not optimum. For monitoring purposes it may be enough to connect the speaker through a sufficiently high resistance to make its loading negligible.

11.29. SENSITIVITY

Unavoidable noise generated within the receiver is not the limiting factor in the sensitivity of most sound broadcast types, which can therefore be measured simply as the number of microvolts (with frame aerials, microvolts per metre) from the generator required to produce standard output. The greater the sensitivity, the smaller the number, which is rather unfortunate as it tends to ambiguity of statement unless when mentioning that the sensitivity is, say, greater, it is made clear whether it is sensitivity in its verbal or its numerical sense that is meant. One way of doing this is to describe the sensitivity as *better*, which is understood to mean greater sensitivity and fewer microvolts.

The procedure is simple: the generator, dummy-aerial, and output-meter connections are as described in the preceding section, and the receiver is tuned accurately to the signal. Either the signal level should be low enough not to operate the a.g.c. or the latter should be put out of action without altering the initial sensitivity. Unless otherwise recorded, manual volume and tone controls are set to give maximum gain.

Novices are sometimes perturbed by the thought that the microvoltage actually across aerial and earth terminals is less than that delivered at the output of the generator, the balance being lost in the dummy aerial; and that the net signal depends on the impedance between aerial and earth terminals. But it must be realised that this condition is less artificial than would be the measurement of the microvoltage at the A and E terminals. The receiver designer may, if he likes, use a low-impedance aerial coupling coil which gets only a small proportion of the generator output ; but as it presumably gives a high step-up ratio to the tuned coil the signal at the grid of the first valve may be just as high as if a high-impedance coupling coil were used. For the same reason, receivers must never be compared by connecting their A and E inputs in parallel.

11.30. SIGNAL/NOISE RATIO

The foregoing simple concept of sensitivity suffices only when, under the conditions specified, the amount of self-generated noise (Sec. 14.32) in the output is negligible. If the gain of the receiver is very high, it may even happen that the noise output alone is 50 mW or more, in which case the maximum-gain 50-mW output standard for sensitivity measurement obviously breaks down completely. Therefore any definition of sensitivity capable of general application must specify a signal/noise ratio. For communication, 15 dB has been proposed, and for sound broadcasting 40 dB. What is called the noise-limited input signal is the minimum input at which any chosen value of signal/noise ratio is obtained.

Suppose that for the purpose in view sensitivity is defined as the minimum r.f. input voltage modulated 30% at 400 c/s to give 50 mW output with a signal/noise ratio of 15 dB, which is a 31·6 : 1 power ratio. Therefore the standard noise output in this case is 1·6 mW. According to one method of combined sensitivity that has been proposed,* the volume control would be set to give a noise output of 1·6 mW with no signal, and then sufficient signal provided to give 50 mW, this signal being regarded as specifying the sensitivity. But if the modulation were then switched off, leaving the carrier wave, the noise

* J. M. Pettit in *Proc. I.R.E.*, March 1947.

would in general be greater than 1·6 mW, so the intended signal/noise ratio would not in fact be obtained. But by making successive adjustments of volume control and signal-generator output until the output with modulation on is 50 mW and with it off is 1·6 mW (or whatever conditions may have been laid down), the final signal input can fairly be taken as a measure of the sensitivity. It may be that a.g.c., if any, is brought into action; this is quite in order. To complete the information, the reduction in a.f. gain needed to achieve the desired result is recorded. It can be measured either by a test using an a.f. signal injected between detector and volume control (e.g., at gramophone pickup terminals), or by turning the volume to maximum at the end of the previous test and noting the output. E.g., if 630 mW, the a.f. loss must have been 630/50, or 11 dB. Care must be taken not to exceed the linear range of the receiver; to bring a large power-ratio within it one may have to reduce the pre-detector gain.

Other methods are given in an article on measurement of sensitivity of a.m. communications receivers (*M.I.*, Sept. 1956).

Strictly, the output power meter ought to be square-law, so as to measure the noise on a mean-square basis, but the error due to using the more usual type (which has a response closer to mean values) seems to be generally tolerated. The error due to measuring signal output with noise present, as just described, can also be tolerated when noise is not less than about 15 dB below. But measurement of low noise levels is likely to be complicated by hum output, which gives a meter reading out of all proportion to its audibility (Sec. 11.18), and it may be necessary to use the weighting amplifier (if *all* noise is to be included) or a filter cutting off at about 300 c/s (if hum and fluctuation noise are to be measured separately). If the standard noise output cannot be obtained even at maximum gain, the situation is covered by the previous section, the signal for standard output being read with gain at maximum.

11.31. NOISE FACTOR

At the lower radio frequencies, self-generated noise is not usually the limiting factor in a receiver provided with ample gain; instead, this factor is external noise and interference, which of course depend on the reception site rather than on the receiver. But at very high r.f. the amplification that can usefully be employed is limited by fluctuation noise. Some of this is due to the resistance of the signal source, so for a prescribed signal/noise ratio there is a minimum signal that would be acceptable even if the receiver were noiseless. But some fluctuation noise is inevitably generated in the receiver itself and further reduces the possible sensitivity. This kind of noise, too, is taken account of in signal/noise ratio, but the expression of sensitivity in terms of this ratio depends on bandwidth, modulation depth, etc., and either input voltage or signal/noise ratio must be specified arbitrarily. A more fundamental quantity is noise factor (or noise figure), defined in B.S. 2065 : 1954 as " the ratio of the total mean-square noise output e.m.f. to that part of it which is due to the thermal noise of the source circuit treated as a passive network at 290° K over the frequency range which can be considered to limit the signal channel of the receiver ". In practice it is most easily measured by means of a noise generator (Sec. 4.39), the maker's instructions for which should be followed. Information can also be found in the paper by Bray and Lowry (footnote on p. 391) and in " The Theory and Measurement of Noise Factor ", by R. J. Yates (*M.I.*, July to Nov. 1950). See also *M.I.*, Sept. 1956, from which the following is quoted.

" To test the noise factor of a receiver, connect a noise generator to its aerial terminal, and connect an output power meter to the a.f. output terminals. With the noise generator connected but generating no output, turn the receiver gain to maximum; switch off or disconnect the a.g.c.; and measure the output power due to internal noise. Switch on the noise generator and adjust its output until the a.f. output power from the receiver is doubled. The noise output from the generator is then equal to N_{eq}*. The noise generator, being calibrated in decibels relative to thermal noise, then indicates the noise factor directly.

" This measurement is of great value because the noise factor obtained is independent of gain and bandwidth. And, providing the controls of the receiver are not altered during the test, their setting has little effect on the results of the measurement. The noise factor can therefore be regarded as an absolute measure of sensitivity, whereas signal/noise ratio is obviously a relative one. In addition to the definitions of noise factor that have been given already, it can be regarded as the relationship between the signal/noise ratio at the input of the receiver and the signal/noise ratio at its output. So the final signal/noise ratio can be calculated for any specified conditions by determining the signal/noise ratio at the aerial for the bandwidth of the receiver and multiplying by the noise factor ".

11.32. SELECTIVITY

Selectivity is the ability to discriminate, by frequency-dependent selection, between the desired signal and signals at other frequencies. To measure this correctly it is essential to have more than one signal at the same time, for some of the ingredients of selectivity depend on interaction between the signal to which the receiver is tuned and the interfering signal on a different frequency. But for simplicity, to enable measurement to be made with one signal generator, or for analysis, these effects are sometimes excluded. The procedure is at first the same as for measuring sensitivity: the receiver is tuned accurately to the modulated signal, taking care that a.g.c. is either not operated by it or is put out of action. The output is not very important so long as the noise part of it is insignificant. The carrier frequency is then varied each side of resonance, and the corresponding signal voltages to

* The sum of the noise generated in the receiver and amplified thermal noise, equivalent to a noise input whose level is given by dividing the noise output by the gain of the receiver. Noise factor is taken as the ratio of N_{eq} to thermal noise.

maintain the same output are noted. The results can be expressed as the bandwidth between frequencies at which the signal input is a specified number of dB above that on-tune, but a more informative presentation is as a selectivity curve (Fig. 11.27).*

Most of the selectivity in a superhet receiver is provided by the i.f. amplifier, so the differences between selectivity curves taken with different r.fs are likely to be slight. If there is a selectivity control, curves should be taken for several settings of it—say maximum, minimum, and half-way.

A ratio of, say, 20 dB at $\pm$ 9 kc/s, obtained as just described, would profess to mean that an adjacent-channel signal 20 dB stronger than the wanted signal would compete with it on level terms (which would be intolerable interference), or that the interference from an equally strong adjacent-channel signal would be 20 dB weaker, or that for an interference ratio of 30 dB the adjacent-channel signal would have to be at least 10 dB weaker than the wanted signal; and so on.

In practice, both wanted and interfering signals are present together, and this causes several effects that modify the selectivity as measured by a single generator. One effect, *modulation suppression*, is due to the detector, which responds less to the modulation of both signals when they are together than when they are received separately.† If one is considerably weaker at that point than the other (as interference has to be, for tolerable reception) the response to it is considerably reduced, while the stronger signal is hardly affected at all. The result is an improvement in selectivity, which depends on the relative magnitudes of the signals at the detector, and also on their absolute magnitudes, for the effect is greater when the detector is operated linearly than at the bottom bend, so it is necessary to mention the signal strengths.

Another effect, *cross-modulation*, tends to reduce the selectivity. The earlier stages of a receiver have little preceding selectivity to protect them from strong interference, and if they are appreciably non-linear the modulation of the interference is thereby impressed on the wanted carrier wave, so that even if the interfering signal is completely removed by subsequent tuned circuits its modulation will be present in the output. There is also an effect called *blocking*, which is the reduction in sensitivity to the wanted signal, caused by the presence of a carrier wave on another frequency; so it too tends to reduce the overall selectivity.

If the object is to find out the capabilities of a given receiver, the appropriate test is one devised to indicate overall selectivity under working conditions. If on the other hand the object is to provide design information, it is more informative to measure the various effects separately.

* Measurement of bandwidth is considered in *M.I.*, Dec. 1956.

† " The Mutual Interference of Wireless Signals in Simultaneous Detection ", by E. V. Appleton and D. Boohariwalla. *W.E.*, March 1932.

11.33. TWO-SIGNAL TESTS

The two signal generators are connected as in Fig. 11.24 (remember to use double-impedance dummy aerials and double the desired signal voltages!). Their modulation frequencies should be different, to enable the signals to be distinguished. For measuring overall selectivity, one generator is first set to give the required frequency, input microvoltage, and modulation; and the receiver is tuned to it, the output being adjusted by the volume control to a suitable level—say one-quarter the

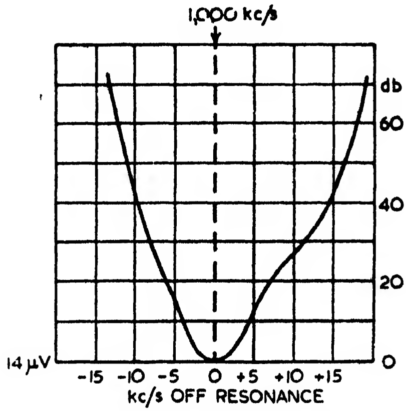


Fig. 11.27—One method of presenting the results of a single-generator selectivity test. Reckoning the input at resonance as zero dB facilitates comparison with other results. The sensitivity is also shown, on the left

rated maximum. The modulation is then switched off, and the second generator set to the required interference frequency and the signal input from it adjusted until its modulation gives an a.f. output reading (representing the tolerable limit of interference) which is the prescribed number of dB below the wanted output previously obtained with the first generator. The ratio of second- to first-generator r.f. outputs is a measure of overall selectivity with the exception of blocking. To determine this, switch off the second generator modulation and switch on the first. Blocking, if appreciable, is indicated by the reduction in a.f. output below its original level. (It is advisable to switch the second generator's carrier off for a moment to check that the original level has not drifted.) Then readjust the volume control and interference alternately until the original wanted level is obtained simultaneously with an interfering carrier wave of the strength which, when modulated, gives the prescribed interference. The ratio of microvoltages is then the overall selectivity ratio.

The selectivity of tuning is measurable separately, as has been described (Fig. 11.27), and blocking has already been measured. The combined effect of modulation suppression and cross-modulation (which work in opposite directions) can be found at the end of the above two-signal test by noting the change in interference output when the unmodulated wanted carrier is switched off. If the interference decreases, the decrease is due to the removal of cross-modulation; if it increases, the increase is due to the removal of modulation suppression. Although this test shows which effect prevails, and the net result of both, it is not easy to find how much there is of each, except under conditions which yield no interference output when the wanted carrier is absent. Then if, when it is switched on, interference is brought in, it must be due wholly to cross-modulation.

It has been assumed so far that the two carrier waves are sufficiently different in frequency for the beat frequency to be removed between detector and output meter. If this is not so, the beat frequency should be removed by a filter; unless, of course, one wants to measure the amount of disturbance caused by adjacent-channel whistle.

In an officially approved test* for mobile a.m. and f.m. receivers, which are required to have an adjacent-channel selectivity of 70 dB at 25 kc/s spacing, the wanted-signal generator is tuned in and set to give a signal/noise ratio of 10 dB. The unwanted-signal generator, tuned to an adjacent channel, is meanwhile giving zero output. Its signal is then brought up until its interference (regarded as noise) reduces the signal/noise ratio to 7 dB. The adjacent-channel selectivity is then given by the signal-strength ratio of the two generators. For this rather stringent test, the generators must have excellent short-term frequency stability and absence of unwanted output signal components. The development of suitable instruments is described by J. F. Golding (just cited) and also by L. R. Head (*M.I.*, Dec. 1958).

11.34. R.F. MODULATION HUM

When measuring small amounts of interference, care must be taken that the output readings are not being affected by hum, which is apt to show up on a meter even when it is not noticeably audible. If necessary, this too may have to be removed by a filter cutting off below about 300 c/s, in which case it may be advisable to use a higher m.f. than 400 c/s.

The measurement of hum itself is described in Sec. 11.18, but faulty design may give rise to an increase when a carrier wave is present, due to modulation of the carrier at hum frequency. So when testing a whole receiver the hum output should be measured with an unmodulated carrier wave tuned in, at say 10 μV, 100 μV, and so on up to the maximum available. It should be noted that this effect may be greater at some radio frequencies than others.

11.35. CONTINUOUS SELECTIVITY-CURVE TRACING

When investigating or aligning selective circuits, especially those intended to give a broad flat top and steep sides, the plotting of

* "Mobile-Receiver Alignment Equipment", by J. F. Golding.; *W.W.*, Feb. 1960.

Fig. 11.28—How selectivity curves appear on the c.r.t. screen when (a) *the r.f. or i.f. signal, and* (b) *the rectified and filtered output are used for Y deflection*

selectivity curves such as Fig. 11.27 becomes intolerably tedious. What is needed is an instantaneous display of the curve, so that the results of adjustments can be seen at once. The apparatus has been described in Secs. 4.36 and 5.38. It consists of a signal generator continuously swept over the required band of frequency, synchronously with the time base of an oscilloscope. The input connection, between signal generator and receiver, is perfectly standard, but the output connection depends on the type of receiver. Although it is sometimes possible to use the final i.f. output, which produces a filled-in trace such as Fig. 11.28*a*, the connection is liable to produce de-tuning or r.f. feedback. If the amplitude is not enough for a reasonable size of diagram, wide-band amplification is needed; at 470-kc/s i.f. this is possible, but not always convenient; at television i.fs it is generally impracticable. The more usual connection is from the detector output, which (if the carrier filtration is effective) yields a line trace as in Fig. 11.28*b*. If the detector carrier filter is not effective enough to give a clear line, it may be necessary to supplement it by some external low-pass filtration; but care must be taken that this does not appreciably affect the shape of the trace. The sweep frequency must in any case be quite low, not only to get through the filter, but to avoid distortion of the curve when the r.f. circuits are highly selective. About 10 c/s is suitable in most circumstances, but receivers with crystal filters may have to be traced out at 1 c/s or even less, using a long-afterglow c.r. tube. As a low sweep frequency may allow time for a.g.c. bias to fluctuate, it may be necessary to put the a.g.c. out of action.

Unless exceptionally high-level detection is a feature of the receiver, a stage of amplification at z.f. or very low a.f. will be needed between the detector and the Y plates.

The use of wobbulators is expounded in detail by R. Brown (*W.W.*, Feb., March and May 1961), including (in the March issue) information on differential and derivative techniques, which greatly extend the possibilities.

11.36. SPURIOUS RESPONSES

As any treatise on the superheterodyne explains, there are several kinds of undesired responses to which it is liable. These, which should be measured as part of any complete test of this type of receiver, can be placed in the following categories:

Image frequency (sometimes called second-channel). This is measured as the image-frequency rejection ratio, which is the number of times greater a signal has to be at the image frequency than at the proper frequency to produce a given output. Suppose the sensitivity of a receiver at 1,000 kc/s is 25 μV, and the i.f. is 470 kc/s. The oscillator will then be working at 1,470 kc/s, the image frequency being $1{,}470 + 470 = 1{,}940$ kc/s. If now the signal generator is adjusted to this frequency, without altering the receiver, and the sensitivity to it is 60,000 μV, the rejection ratio is $60{,}000/25 = 2{,}400$, or nearly 68 dB. The higher the frequency, the lower the ratio is likely to be. Some broadcast receivers provide very little protection indeed against image interference on short waves, and everything comes in at two positions on the tuning scale. It is usual to measure image ratio at the same time as sensitivity and to plot both against carrier frequency.

Intermediate frequency. The same procedure is followed for measuring the i.f. rejection ratio, except that the generator is tuned to the receiver's intermediate frequency for the second reading. The ratio is likely to be least when the receiver is tuned to the frequencies nearest the i.f.

I.f. harmonics. When the i.f. signal reaches the detector it is distorted thereby and strong harmonics are produced. If these can work round to the receiver input when it is tuned to those frequencies, they will interfere. They should therefore be looked for by tuning the receiver to a signal at multiples of the i.f.—e.g., 940 and 1,410 kc/s if the i.f. is 470 kc/s.

Other undesired responses. There are many other conditions that may give whistles, some requiring only one incoming signal for their formation and others two, and complete data on them involve a large amount of work. Detailed instructions are given in the paper by Bray and Lowry (reference at foot of p. 391).

11.37. AUTOMATIC GAIN CONTROL

To plot a.g.c. action it is desirable that the output stage should not be overloaded, so it is necessary to use the volume control to restrict the output to, say, a quarter of the rated maximum. The method of beginning with the strongest input (1 V), controlled to give this output power, and then reducing it, is subject to the objection that it entirely fails to give an important item of information—the output level to which the a.g.c. tends to bring signals that overcome the initial bias. If that level is too high, a large proportion of the volume-control range, corresponding to gross overloading, is wasted, and the remainder is consequently overcrowded. Also the method involves measurement of very small output powers towards the lower part of the a.g.c. curve. And it is not possible to conform universally to the standard specified, because many signal generators are incapable of an output of 1 V.

The author's procedure avoids all these disadvantages. The volume control is set at its maximum. If with no signal there is appreciable receiver noise output, its amount is noted. Then a standard signal of increasing strength is applied and the corresponding output observed. A convenient series of points is 1, 3, 10, 30, 100, etc., mW; 3 is near enough to $\sqrt{10}$ to be placed half-way between 1 and 10 on a logarithmic scale without serious error. When a stage is reached such that the next step would cause the receiver output to exceed about a quarter of the rated maximum, the volume control is adjusted until the output is reduced to one-tenth (in watts), and the next two output meter readings are multiplied by ten. If necessary this process is repeated, until the maximum generator signal is reached.

Table 11.1

Output measured (mW)	*Output plotted (mW)*	*Input (µV)*
2	2	0
3	3	0·6
10	10	1·3
30	30	2·5
100	100	4·2
300 } * 30 }	300	7·7
100	1,000	12·5
300 } * 30 }	3,000	44
100	10,000	1,130
300 } * 30 }	30,000	33,000
100	100,000	400,000

* Volume control adjusted

Table 11.1 is an example (Fig. 11.29*a*) for a receiver with a nominal output of 3½ W.

The a.g.c. curve is plotted as *extrapolated output watts* against r.f. microvolts, and shows the output that would be given if the audio stages were unlimited in power-handling capacity. A receiver with an excessive delay voltage or too much audio gain can produce theoretical kilowatts in this way! Faulty design is thus revealed.

If a briefer test is required, the extrapolated output watts at 10^4 µV is perhaps the most useful single datum. Experience shows that with 30% modulation this should normally be slightly greater than the rated maximum output. In Fig. 11.29*a*, for example, it is excessive.

When a muting system is incorporated, the lower reaches of the curve are especially interesting. If an appreciable range of signal strength is occupied in overcoming the suppression, so that the first

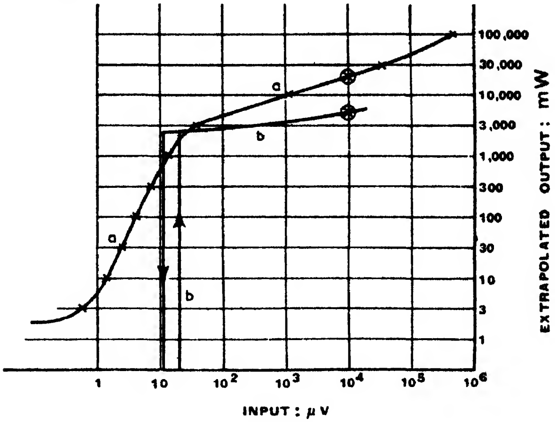


Fig. 11.29—Example of a.g.c. characteristics. Curve a *is taken with no muting system in operation. At the extreme left a noise output of* 2 *mW with no signal is shown; then there is a steep rise during which a.g.c. is inoperative because the bias voltage has not been overcome. When it is, a.g.c. restricts further rise. Owing to excessive bias, the flat top is too high. With muting in use, curve* b *results*

part of the curve is similar to that of ordinary biased a.g.c. but steeper, one result is more-or-less noticeable distortion of signals in this region of strength. To make sure that this does not take place, designers usually try to obtain a slight backlash effect, so that by however small a margin the suppression bias is exceeded, it is completely thrown off with a jerk (take care of the output meter!) and the signal can be carried appreciably below this critical value before suppression again takes charge. The increasing and decreasing characteristics should be distinguished by arrows (Fig. 11.29*b*).

11.38. OVERALL DISTORTION

Although most of the distortion of the a.f. output from a sound receiver is usually due to the a.f. amplifier, and the testing of that has been covered in Secs. 11.8 to 11.19, the preceding stages are liable to cause additional distortion in various ways, and complete receiver tests would certainly include data on overall distortion. In a vision channel these stages may account for most of the distortion.

The methods are fairly obvious extensions of those for a.f. and v.f. amplifiers respectively; the fact that between the a.f. source and the a.f. amplifier there are stages where the original signal form exists only as the modulation of a carrier wave need not complicate the procedure very much. It is necessary, however, to consider whether a.g.c. would upset the results, and if so to take appropriate action.

Excessive selectivity and unsuitable detector constants may considerably reduce the bandwidth provided by the m.f. (a.f. or v.f.) amplifier alone. Fig. 11.30 shows typical overall curves, which are usually plotted with the response at 400 c/s (or the maximum response) as zero-dB level. The signal generator is modulated by an oscillator of variable frequency, and the output noted over the whole m.f. band. The signal strength, depth of modulation, and output power, within reasonable limits, have little or no bearing on the results, and standard conditions can conveniently be assumed. On the other hand, carrier-wave frequency may have a pronounced effect; the settings of tone

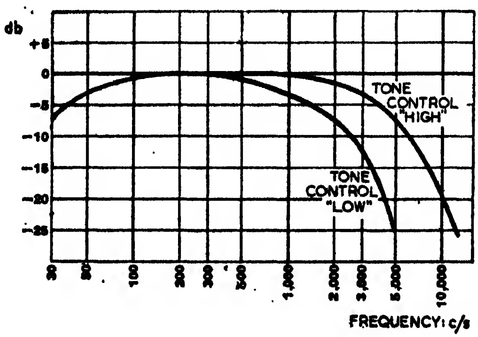


Fig. 11.30—Typical overall modulation-frequency response curve of a receiver

control and selectivity control obviously have, and separate tests are necessary to show them.

The need for intelligent interpretation of results must be re-emphasized in connection with these frequency-characteristic curves. The loudspeaker characteristics may alter the overall result very greatly. A severe falling-off in response at the upper frequencies may be largely filled up by speaker resonance. The impedance of the speaker rises, too, and with a pentode output stage this tends to increase acoustical output, whereas with a triode the opposite obtains. If this is fully kept in mind, however, frequency characteristics can be valuably informative.

Non-linearity, although affected—perhaps considerably—by the modulation frequency, depends mainly on signal amplitude. It can be examined in any of the ways already described. One can obtain a dynamic transfer characteristic by comparing (on the c.r. tube) the modulating signal in the generator with the output from the receiver. If distortion-measuring equipment is available, the non-linearity can be measured as harmonics or intermodulation and plotted against m.f. output power. Note that, on account of the detector, non-linearity increases at low signal levels as well as at high. Although it does not show up all possible non-linearity, a static test can be done with less equipment and trouble and gives a good idea of the overload point of the receiver. A signal modulated 5 or 10% is adjusted to give an output comfortably within the capabilities of the receiver.

The modulation depth is then increased in suitable steps to the maximum possible, and the output watts are plotted against the square of the modulation percentage. In a perfect amplifier the result is a straight line passing through the origin. The output power at which a serious departure from perfection is made can be seen. Of course the modulation system of the generator must be above reproach, or the test will be vitiated. To test it, the whole modulation range should be run through *within* the capabilities of an amplifier. Unless it is desired to include possible a.g.c. effects it may be desirable to cut out the a.g.c. action, as some systems are affected by the depth of modulation.

11.39. TUNING DRIFT

The extent to which receiver tuning shifts with temperature, mains or battery voltage, signal amplitude, etc., mainly as a result of variations in the frequency-changer oscillator, is important. (Incidentally, such changes may cause errors in other measurements if not noticed. To minimize temperature variations the set should be run for not less than about 15 minutes before testing.) Tuning drift is observed by using the receiver's beat-frequency oscillator, if any, or a constant external i.f. oscillator coupled to the i.f. amplifier, to produce an audible beat frequency with an incoming constant-frequency signal, derived from a crystal-controlled oscillator or broadcasting station. To measure the drift precisely in c/s this beat note may be compared with a calibrated a.f. oscillator, starting from zero.

When plotting drift against time after switching on from cold, it is usual to allow a period of from 1 to 5 minutes—the actual amount must be recorded—before readings are taken. As the drift is usually about the same percentage of oscillator frequency at all frequencies, in c/s it is likely to be roughly proportional to the frequency to which the set is tuned; but this is not necessarily so if compensation is used. During this test, supply voltage and other conditions should be kept constant, and the ambient temperature should be noted.

Frequency change may be plotted against mains voltage over the range of voltages at which the set is rated to work on any one mains-voltage tapping, or against the range of voltage in the working life of the battery.

Any change resulting from variation of signal input, from the lowest needed to give standard output up to the highest the generator can give, should be measured. It is obviously necessary to make sure that the generator frequency itself is not affected by adjustment of its attenuator.

11.40. VISION-CHANNEL TESTING

Although testing of the vision channel has been included (either directly or by implication) in the foregoing sections, a summary may be useful.

Measurement of sensitivity is similar to that for sound, except that

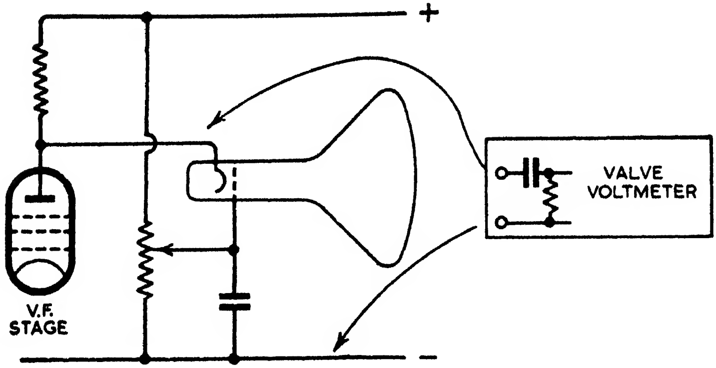


Fig. 11.31—Method of connecting a valve voltmeter to read the output of a vision channel. The voltmeter must be provided with a low-capacitance probe

the output is read as voltage at the control electrode of the c.r.t. Fig. 11.31 shows the conventional arrangement of this part of the receiver, and where to connect the valve voltmeter. Note that there is a z.f. potential to be excluded; if the circuit is live to mains this must be borne in mind. The receiver should be connected with the live wire of the mains to +, and even then the voltmeter ought to be well insulated from earth. The amount of connection capacitance that can be tolerated depends on the modulation frequency used: if it is low, say 1 kc/s, no special care is needed; in fact even at 10 kc/s about 50 pF can be tolerated, but if the full v.f. band is to be measured it is necessary to restrict stray to about 2 pF, which calls for a special probe (Sec. 5.13).

The most important class of test is for assessing distortion of the v.f. waveform. As with a.f., there are two ways of tackling this. One is to obtain a frequency characteristic, which ideally should be flat from 25 c/s to about 3 Mc/s (for the British system). While observed departures from this give a good deal of information both on the resulting shortcomings of the picture and on the faults in design or adjustment causing them, more and better information can be obtained by pulse or square-wave testing.

A curve of overall response against v.f. necessitates a signal generator capable of being modulated over the whole v.f. band, at least from 25 c/s to 3 Mc/s, and an output voltmeter with the low-capacitance input just mentioned. Plotting the curve takes some time, because even a slight resonance before cut-off causes sufficient transient distortion to affect the picture. For adjusting receiver circuits, this procedure is far too slow. Continuous curve-tracing equipment is possible, but rather elaborate, and the sweep speed is restricted. Since faulty adjustment is more likely to occur in the pre-v.f. circuits, and wobbulator equipment is simpler and can be run faster, it is more often used. The carrier wave, unmodulated in amplitude, is swept over a band up to about 10 Mc/s wide covering the channel to which the receiver is tuned. A convenient sweep speed is 50 c/s. The output of the detector is amplified and applied to the Y plate; the amplifier should be effective from a few c/s up to several hundreds. Various arrangements have been devised to provide a frequency scale; one is a tunable absorption frequency meter that causes a dip in the trace at the frequency to which it is set; another is a calibration oscillator producing a " pip " at every Mc/s. A wobbulator is useful for checking the decoupling of a vision i.f. amplifier. If the receiver chassis is not live, this can be done quite simply by putting a finger on a supposedly decoupled point and seeing if any change occurs in the displayed response curve.

Although complete frequency-response *and phase* data theoretically

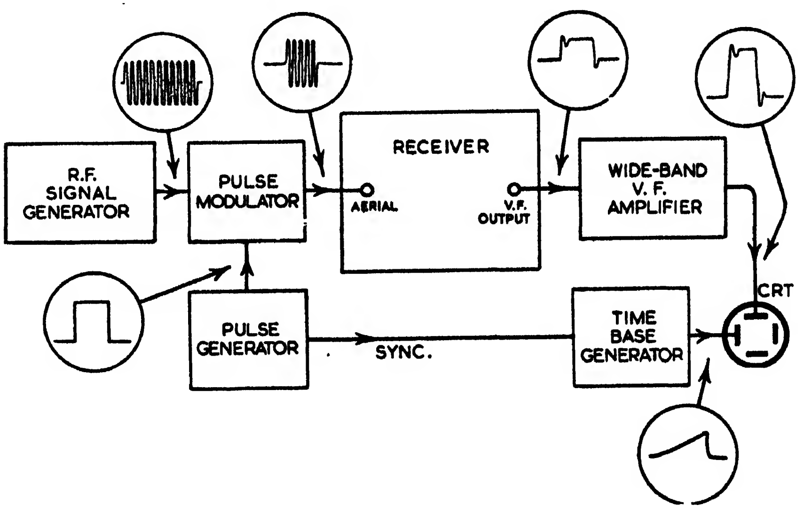


Fig. 11.32—Block diagram of apparatus for overall pulse testing of a television receiver

are sufficient to enable the time response to any shape of signal to be predicted, for television purposes it is more generally helpful to use a time-response form of test, as already considered in Secs. 11.19 and 11.21, except that of course for overall testing the pulses are used to modulate a r.f. signal generator. Fig. 11.32 shows the general arrangement. The whole of the signal-generating equipment must be designed to provide a close approach to a perfect pulse, say 10 μsec duration with a rise time not longer than 0·02 μsec and negligible overshoot or sag. Better still is the pulse-and-bar signal of Sec. 11.21. At the output end the connection should impose a capacitance not exceeding 2 pF and be followed by an amplifier with an output up to about 50 V over a band up to at least 5 Mc/s and preferably more. A description

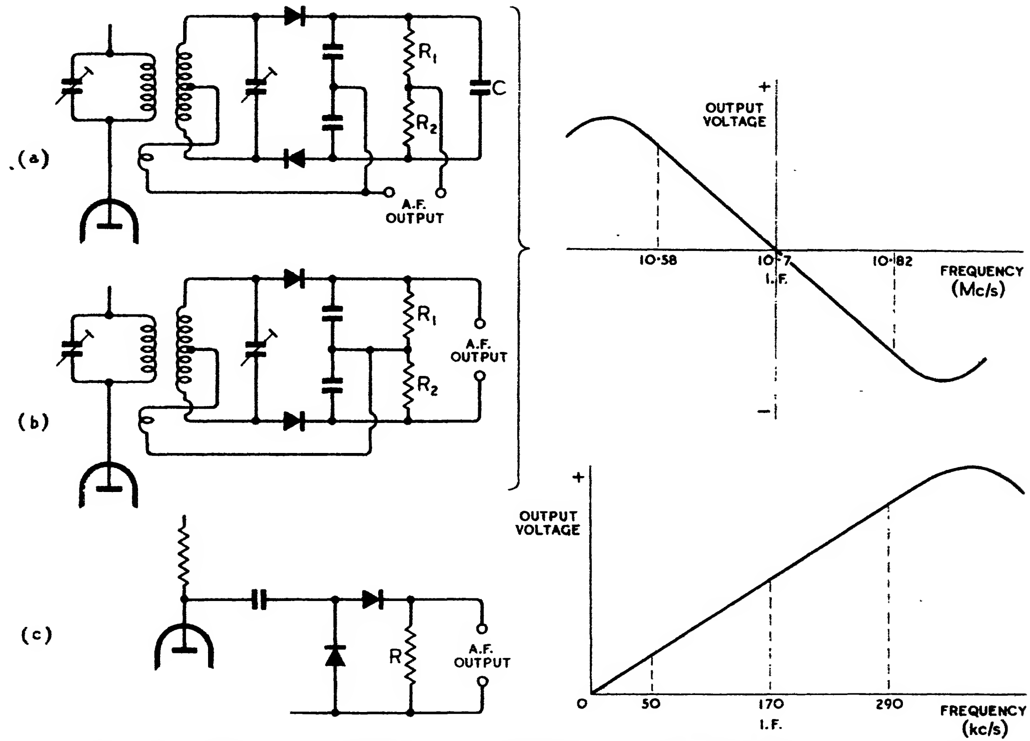


Fig. 11.33—Most f.m. discriminators are included among these three types: (a) *ratio,* (b) *Foster-Seeley, and* (c) *pulse-counter. Typical corresponding characteristic curves are shown on the right*

of such equipment, and a discussion of results, is given by M. V. Callendar (*W.W.*, Feb. 1952).

11.41. F.M. RECEIVERS

Although the preceding Sections on receiver measurements refer particularly to amplitude-modulated receivers, most of the instructions apply also to frequency-modulated receivers. The differences will now be considered. They arise mainly, of course, from the different function of the detector, which embodies a frequency discriminator.

Then, although there are such things as narrow-band f.m. receivers, better noise suppression is obtained with wide frequency deviation, which necessitates a correspondingly wide acceptance band. The standard frequency deviation for broadcast receivers is 75 kc/s each side of the nominal carrier frequency, and for communication receivers 15 kc/s. The sidebands extend somewhat wider, and for $\pm$ 75 kc/s deviation a total band or channel of 240 kc/s may be assumed. This is too much for medium or high frequencies, so f.m. necessitates v.h.f. —for broadcasting, 87·5 to 95 to 100 Mc/s, with 10·7 Mc/s i.f.

Another characteristic of f.m. that must be allowed for in receiver testing is the pre-emphasis of high a.f. in broadcast emissions.

11.42. THE DISCRIMINATOR

The commonly used ratio discriminator has an inherent limiting action, so a preceding limiter, though desirable (to reduce liability to multipath distortion), is often omitted. The Foster-Seeley discriminator differs more in principle than the close similarity of circuit (Fig. 11.33) might suggest, and a separate limiter is essential. The pulse-counter discriminator* is quite different both in circuit and principle, and requires a limiter. The main purpose of Fig. 11.33 is to show where instruments should be connected to measure the a.f. and/or z.f. outputs. In (a) the diodes are connected so that with an unmodulated carrier they give nominally equal z.f. potentials across R_1 and R_2 which add, and the total is kept free from a.f. by the large capacitor C. The a.f. output is developed between the terminals so marked. In (b) the diodes are connected to give z.f. potentials across R_1 and R_2 in opposition, so again the a.f. output is free from z.f., except any due to imperfect balance. In (c) both z.f. and a.f. are developed across R.

One of the most important characteristics of a f.m. receiver is linearity of the discriminator over the whole channel bandwidth. Typical correct examples are shown in Fig. 11.33, the central or unmodulated frequency in each case being the i.f. An input signal of this frequency, variable to the extent of 150 kc/s or so each side, and of sufficient amplitude to be effectively limited, is applied to the grid of the last i.f. or limiter stage.

In the simplest method it is varied by hand and the output is plotted point by point from the readings of a valve voltmeter connected across

* See " Low-distortion F.M. Discriminator ", *W.W.*, April 1956.

the a.f. terminals. Disadvantages of this method are that it is tedious, and that being a static test it does not necessarily give exactly the same results as under working conditions. In the case of the ratio detector (a) it certainly would not unless the voltage across C were artificially maintained constant. This can be done by measuring the voltage at the frequency giving zero output, and connecting a low-resistance source of equal voltage across it.

Adequacy of input can be ensured by a preliminary test in which the signal is increased well beyond the point at which the z.f. output ceases to rise steeply.

A quicker and truer-to-life method is to use a wobbulator, substantially as described in Sec. 11.35 for a.m. receivers. But the input and output connecting points are as just specified, and there is of course no need to stabilize the voltage across C in Fig. 11.33(a). Instead of the comparatively low sweep frequency recommended for a.m. receivers, the frequency should be within the a.f. band; in fact, if the apparatus permits, several representative a.f. frequencies should be used, to detect any dependence of discriminator characteristic on modulation frequency. For this work a f.m. signal generator with provision for modulation from an external source, and an oscilloscope with the X voltage accessible for that purpose, are likely to be more useful than the usual a.m. wobbulator equipment which may be limited as regards sweep frequency. The most important thing, however, is linearity of the oscilloscope.

11.43. SENSITIVITY

In a.m. receivers the a.f. output power is, or should be, linearly proportionate to the square of the r.f. input voltage, up to the maximum " undistorted " output—apart from the effects of a.g.c., which would be put out of action for such a test. So the choice of 50 mW as standard output is fairly arbitrary. But a f.m. receiver does not achieve its main purpose—noise and interference suppression—unless the signal is sufficient to bring the limiter into action. The least input of which this is true is therefore a measure of the effective sensitivity of the receiver.

It can be found by connecting a v.h.f. signal generator to the aerial socket of the receiver, and injecting several millivolts of unmodulated signal. Even with the volume control turned well up the noise level should then be very small. Next, the input is reduced until a sudden increase in noise takes place, and the signal microvoltage at the noise rise point is noted.

11.44. SIGNAL/NOISE RATIO

The definition of sensitivity as the minimum modulated input to achieve a prescribed signal/noise ratio, applicable to sensitive a.m. receivers, is not normally appropriate to f.m., for the reason given in the previous Section. But it may sometimes be interesting to measure the signal/noise ratio, above or below the limiting level. The simplest method is to inject a signal having the frequency, amplitude and modulation at which the s/n ratio is required, and measure the a.f. output power. Call it P_1. The modulation is then switched off, leaving only noise. The reading is then P_2, and the s/n ratio is $(P_1 - P_2)/P_2$. If the signal is above limiting level a very sensitive output meter will be needed to measure the noise.

Ideally the output meter ought to read mean-square values, but such an instrument is seldom available. The error due to using the usual mean-voltage-squared types would be negligible above limiting level, but might be substantial at small s/n ratios.

11.45. DISTORTION

Non-linearity of the discriminator curves obtained as in Sec. 11.42 can be analysed in the ways described from Sec. 11.13 onwards. Additional non-linearity is liable to be caused by excessive peakiness or inadequate bandwidth of the r.f. and i.f. tuning, especially if the limiting is poor or the signal level low. Such distortion can be determined by moving the generator connection from the limiter grid to an appropriate point nearer to or at the aerial socket. Distortion due to a.f. circuits is similarly included by moving the indicator connection from the discriminator output to the loudspeaker terminals.

When observing the overall a.f. amplitude/frequency characteristics of f.m. receivers there is a complication owing to the de-emphasis circuit immediately following the detector. This is included to offset the pre-emphasis at the transmitter, thereby effecting a useful reduction in noise. There are difficulties in pre-emphasizing the upper frequencies to the required extent at the signal generator, and it is simpler to augment the output readings at the upper frequencies in accordance with Table 11.2, which is appropriate to the standard 50-μsec de-emphasis used for broadcast reception.

Table 11.2

Frequency, kc/s	*dB to add*	*Frequency, kc/s*	*dB to add*
0·50	0·11	5	5·40
0·75	0·24	6	6·58
1·00	0·41	7	7·66
1·25	0·63	8	8·65
1·50	0·87	9	9·54
1·75	1·14	10	10·36
2·0	1·44	12	11·83
2·5	2·09	14	13·09
3	2·76	16	14·20
4	4·12	20	16·07

CHAPTER 12

Very High Frequencies

12.1. BOUNDS OF V.H.F.

ALTHOUGH references to the suitability or unsuitability of methods or apparatus at very high frequencies have been included in the preceding chapters, some notes devoted specifically to such work may be helpful to readers who are unaccustomed to it. Generally there is no sharply defined frequency at which a particular form of technique becomes inapplicable. It is necessary to use judgment based on common sense, experience, and estimation of the relative quantities affected by frequency. Effects negligible at, say, 1 Mc/s become appreciable in accurate work at 10 Mc/s, and dominant at 100 Mc/s. Most of the methods described in this book for radio frequencies present no special difficulty due to frequency as far as one or two Mc/s, and with perhaps a little extra care can be used satisfactorily up to 10–20 Mc/s. Above 30 Mc/s one enters the field of " very high " frequencies (defined as those between 30 and 300 Mc/s, corresponding to wavelengths between 10 metres and 1 metre) and in it a fairly large number of modifications have to be made to the ordinary technique. Above 300 Mc/s still more drastic changes in method are necessary. Such technique requires a book to itself; this chapter is confined mainly to v.h.f.

12.2. HOW CIRCUIT DIAGRAMS CAN MISLEAD

One thing that the experimenter has to get used to, even at moderately high frequencies, is not thinking too much in terms of circuit diagrams, with their localized inductances and capacitances. Whereas the quantities not shown on the diagram are " strays " at moderate radio frequencies, they often usurp control at very high frequencies and become more important in the operation of the circuit than the legitimate components. For example, the action of the oscillator circuit of Fig. 12.1*a* is not at all clear if account is taken only of the components shown., But when the interelectrode capacitances of the valve are marked in as capacitors, the circuit is immediately recognizable as the Colpitts (Fig. 12.1*b*). Incidentally, this circuit is quite a good one for general use in the v.h.f. band.

Capacitors themselves must be recognized as series resonant circuits, with inductance that is appreciable at v.h.f., even when they are classed as non-inductive. So selecting by-pass capacitors is not simply a matter of the more capacitance the better; what is wanted is the least possible impedance, and if more capacitance is used than is needed to tune its own inductance to the frequency to be by-passed, the impedance is greater than it need be. As the author showed (*W.W.*, 29th Sept. 1933) there is an optimum capacitance for every frequency. The inductance of most present-day capacitors is about equal to that of a piece of wire joining the terminal points; assuming this distance to be

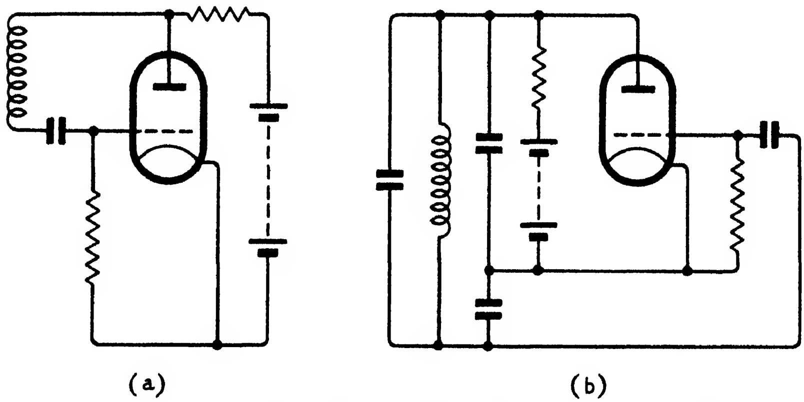


Fig. 12.1—The oscillator circuit which, according to the components actually visible, would be drawn as at (a), *is better understood when valve interelectrode capacitances are shown* (b)

about an inch the optimum capacitance is given in Fig. 12.2. This is confirmed in greater detail by R. Davidson (*W.W.*, Aug. 1952). But the less the inductance can be made to be, the greater the optimum capacitance and the more effective the by-pass over a wide range of frequency. Modern ceramic disk capacitors are produced with this in mind. And the short-circuit effected by a capacitor connected as in Fig. 12.3*a* can be improved by connecting as at *b*. In bush capacitors, which can be bought, or made up as required with metal and mica washers (Fig. 12.4), the inductance is only a small fraction of that assumed in Fig. 12.2. On the other hand, most coils (except the very small ones used for tuning) behave as capacitors at v.h.f., since they are used above their resonant frequency.

12.3. IMPEDANCE LIMITATIONS

The fact that every inch of wire represents an appreciable inductance means that it is difficult to get a very low impedance between two separated points, and this necessitates much thought when laying out and wiring v.h.f. equipment. At the same time, it is difficult to achieve a very high impedance anywhere, because of the stray capacitance in parallel with everything.

Because the distinction between inductance and capacitance is so blurred, it is often easier to accept that fact completely and, instead of

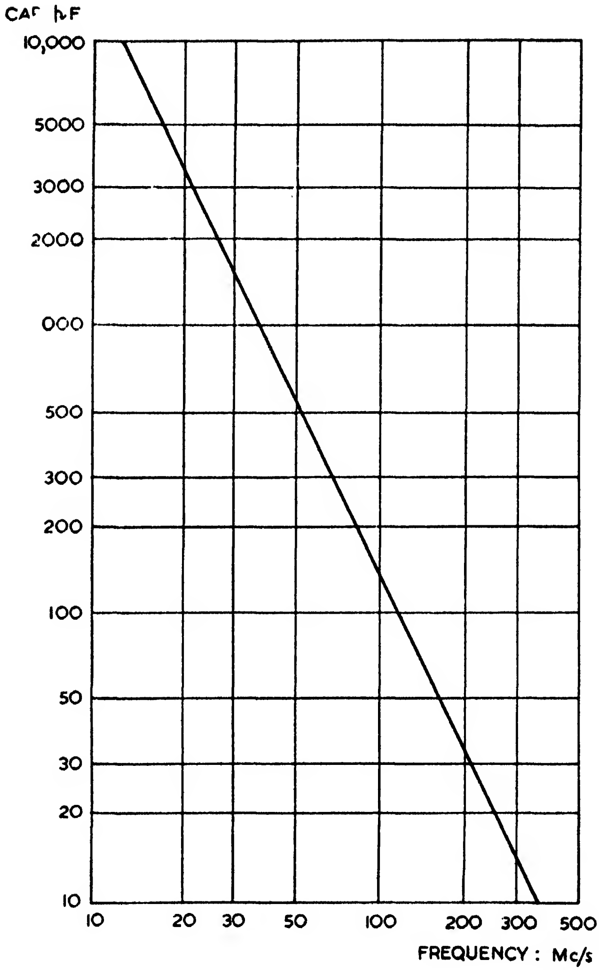


Fig. 12.2—Optimum capacitance of by-pass capacitors, assuming the total length of connections (including the capacitor itself) is one inch

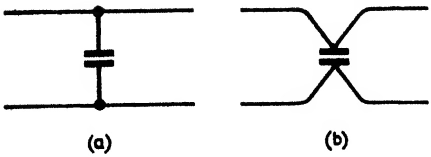


Fig. 12.3—When a by-pass capacitor is connected as at (a) *its impedance is increased by the common connecting leads; this is obviated by connecting as at* (b)

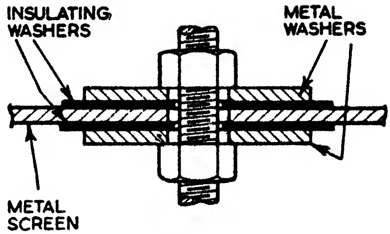


Fig. 12.4—Section of low-inductance by-pass to screen, formed by bushing a lead passing through

attempting to continue with the lumped-component technique suitable for lower frequencies, to use completely distributed reactance of simple forms such as parallel wires or tubes, or coaxial cylinders, whose electrical characteristics can easily be calculated from their dimensions (Sec. 14.27). A required amount of reactance, or a complete resonant circuit, can be provided in this way, usually with less calculation and higher Q than any lumped combination. The resonant frequency is known in terms of the corresponding wavelength; e.g., a parallel or coaxial line quarter of a wavelength long and short-circuited at one end forms a simple resonant system. There is no need to calculate inductance or capacitance or have doubts about the frequency; the only measuring instrument needed is a centimetre scale. And it is usually possible to dodge the difficulties due to the impedance of connecting leads by making lead and tuned circuits one and the same.

12.4. VALVES AT V.H.F.

At some frequency in the v.h.f. region the behaviour of conventional types of valves begins to alter rapidly with frequency. There are various causes, but they add up to a drastic reduction in input impedance. An increasing capacitive admittance, proportional to the capacitance of the control grid to other electrodes, is only to be expected; but a less obvious effect is a great increase in conductance. Whereas at low or even moderately high radio frequencies the input resistance of a negatively biased valve is for practical purposes almost infinite, at v.h.f. it is usually only a few thousands of ohms (Fig. 13.6). This is partly the result of a phase shift due to transit time—the time taken by the electrons to cross the interelectrode space—and partly to the inductance between the electrodes and the nearest points at which connections can be made to them. When the input resistance falls below a certain amount, the valve is no longer capable of amplifying—and therefore of oscillating—no matter how well-designed the circuits.

To meet the demand for television, f.m. broadcasting, and v.h.f. communication, the frequencies at which valves continue to be effective have been pushed higher and higher. But to derive full benefit from their characteristics it is necessary to design the circuits to fit. The makers' data for valves suitable for v.h.f. include the input resistance at a specified frequency, and the resistance at other frequencies can be estimated on the assumption that it is inversely proportional to frequency squared. If, for example, the input resistance is 5 kΩ, there is no point in striving to provide coupling circuits with a dynamic resistance many times greater.

12.5. NOISE

Man-made noise and atmospherics decline with increasing frequency, and the limiting factor becomes what is called fluctuation noise (Sec. 14.32), caused by random movement of electrons in circuits and valves. The power of such noise is proportional to the absolute temperature of the circuit, and the accepted frequency band (*W.W.*, May 1956, p. 235, and June, p. 266). In valves it is proportional to anode current, but also depends on various features of valve design; e.g., it is greater in multi-electrode valves than in triodes (*W.W.*, Dec. 1960, p. 623). For that reason, cascode-connected double triodes are preferred to pentodes as first-stage valves.

The object in v.h.f. design is to increase signal/noise ratio, so a change that would increase noise could be justified if it increased signal more. Obviously the most important noise producers are the resistance at the input of an amplifier or receiver, and the first valve, because these are followed by the greatest amplification. In general, a frequency changer contributes several times as much noise as a straight amplifier. The amount of noise contributed by a valve working under specified conditions can be expressed as the input resistance which would produce the same amount of noise if the valve itself were noiseless; this is the *equivalent noise resistance* which is included among the valve data. The *noise factor* (Sec. 11.31) of the first stage is the most significant information about a v.h.f. receiver, because it discloses its potential sensitivity. It can be defined as the ratio of the output and input signal/noise ratios, and is usually expressed in dB.

Where the greatest possible sensitivity is needed (e.g., radio telescopes and some kinds of radar) use is made of masers and mavars (often called, not very wisely, parametric amplifiers). As they are rather specialized types of equipment, and are most often used at still higher frequencies than v.h.f., they are not included here, but for general information see *W.W.*, April and May 1959.

12.6. OSCILLATORS

As a simple example of how the use of ordinary valves and components should be modified at a moderate frequency in this band, consider an oscillator to work at, say, about 90 Mc/s. To get the close coupling necessary to ensure oscillation when the impedance of the circuit is unavoidably so low and is further lowered by the shunting effect of the valve, and to make the system as rigid as possible so as to prevent vibration affecting the frequency of oscillation, single-coil circuits such as the Hartley or Colpitts are preferred to coupled-coil arrangements. To ensure ready oscillation over the whole of the band, the effective position of the tapping must be considered. At lower frequencies the cathode can be tapped near one end (usually the grid end) of the coil with complete success. But at v.h.f. the tap should be near the centre. If stray capacitance comes across a portion of the

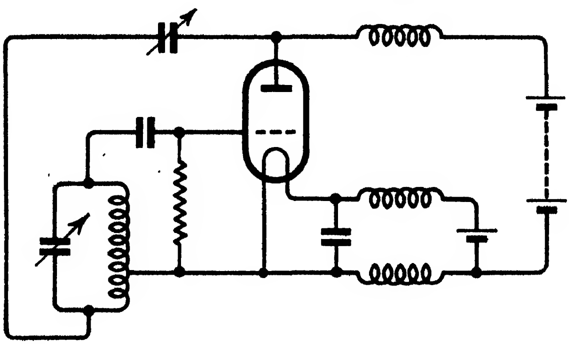


Fig. 12.5—Typical oscillator circuit for v.h.f.

coil the effective tapping-point may thereby be shifted along to an unfavourable position.

For rigidity and low capacitance the coil should be wound of thick bare wire—say about 16 s.w.g.—shaped on a rod $\frac{3}{8}$ in. in diameter and mounted as directly as possible to the valve holder, which should be of special low-capacitance shape and material and mounted an inch or so clear of the baseboard. The variable capacitor, also of compact design and with maximum and minimum capacitances of perhaps 25 and 3 pF respectively, is mounted so as to make wiring practically non-existent, and especially so as to avoid what are, in effect, unauthorized turns or half-turns of coil; and of course it should be rotated by a slow-motion control at the other side of an earthed screen. The grid-leak resistance can be quite low—25 kΩ for example—because the input resistance of the valve is in any case low, and with a high resistance there is a risk of squegging.

The oscillation can be controlled by a small trimmer slung in the wiring. Three chokes are shown (Fig. 12.5) for isolating the batteries or power unit in case it is desired to put them at some distance. In any case it is a good thing to keep the r.f. currents in their own quarters. The filament chokes must be of reasonably low resistance, and, assuming a filament consumption of about 0·1 A, they can be made by winding 50 turns of 26 s.w.g. wire on a $\frac{1}{2}$-in tube. Where the resistance

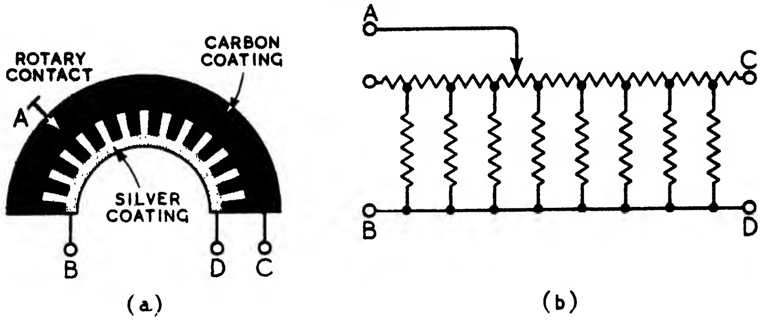


Fig. 12.6—(a) *Rotary attenuator, suitable for v.h.f. It is equivalent to the ladder form shown at* (b)

restriction does not apply, about twice the number of turns of fine wire on a ¼-in former is to be preferred.

Towards the highest frequency in the band (300 Mc/s), it is necessary to use a type of valve designed for v.h.f. For example, the EC71 is a subminiature triode for use as an oscillator up to 500 Mc/s. And there are many types—some comprising a balanced pair of valves for push-pull oscillators—suitable for generating a substantial amount of power up to several hundred Mc/s for standing-wave tests, etc. It is usual for the tuning circuits to be parallel or coaxial lines. Detailed information on such an oscillator covering 150 to 500 Mc/s has been given by J. H. Andreae and P. L. Joyce (*W.W.*, April 1958).

With standard-signal generators the chief difficulty is to get the signal where one wants it and nowhere else. Attenuators of the conventional types become difficult and the waveguide type comes into its own (Sec. 6.22).

But an unconventionally constructed form of the conventional ladder type, described by B. G. Martindill (*W.W.*, April 1957), would seem to maintain its characteristics from zero up to the highest v.h.f. and be simpler and cheaper to produce. The elements consist of carbon coating on a thin sheet of insulating material, suitably stamped out. Fig. 12.6*a* shows one convenient form, which is obviously equivalent to the conventional Fig. 12.6*b* (Secs. 4.32 and 6.21).

12.7. TRANSISTORS AT V.H.F.

Even more than valves, transistors tend to fall off in performance at v.h.f.—or much lower frequencies—and, although methods of manufacture have been devised which give good working characteristics at v.h.f., these methods involve extremely close tolerances, so in general are more costly and would not be justified for general-purpose transistors. In short, special types are required.

One of the main limiting effects is the transit time of carriers (holes or electrons) across the base layer. The conditions are quite different from transit time in valves. In transistors the carriers travel relatively slowly, by diffusion, instead of rapidly under the influence of a strong electric field. This introduces not only delay but dispersion. The results are phase shift and reduction in current gain. The extent of the latter is usually denoted by f_α, which is the frequency at which α is less by 3 dB than at low frequencies (i.e., is reduced in the ratio $\sqrt{2} : 1$. An alternative parameter f_T is the frequency at which $\alpha_e = 1$. Again in contrast to valves, whose transit-time effects occur at frequencies much higher than those at which interelectrode capacitances are significant, transistors are affected at comparatively low frequencies. Since f_α is approximately inversely proportional to the square of the base thickness, the most effective policy in producing transistors for v.h.f. is to make this thickness very small (of the order of millionths of an inch). Another expedient, in the so-called drift transistors, is to graduate the base impurity so as to set up an electric field therein, to speed the carriers on their journey.

Both junctions in a transistor have electrical capacitances, which shunt the emitter and collector resistances in the T equivalent circuit (Sec. 14.22) and introduce phase shifts and feedbacks impossible to summarize briefly. The collector junction capacitance, though the smaller of the two, is the more significant. One effect it can introduce is similar to that of C_{ag} in a triode valve—tendency to oscillate in amplifier circuits. Up to a point it can be reduced by higher collector voltage, which draws the " plates " of the capacitor apart, but usually neutralization is necessary.

12.8. FREQUENCY MEASUREMENT

The heterodyne frequency meters, so useful at medium frequencies, become increasingly unworkable as the frequency is increased. When a change of frequency of 0·0005% sends the beat note completely beyond audibility, the effort to find a beat note, and having found it to hold it, is too great except with relatively elaborate apparatus.* So the absorption wavemeter (Sec. 6.16) is the usual device for finding the frequency. Actual measurement of frequency, or rather wavelength, by means of parallel wires is described in Sec. 6.17. As for receivers, a useful type for v.h.f., by reason of its flat tuning and great sensitivity, is the super-regenerative. The higher the frequency the more workable it is.

Fig. 12.7 shows a simple receiver which, with appropriate tuning inductance, can be used at any frequency in the v.h.f. band. The aerial, if required, is simply a piece of stiff copper wire half a wavelength long, with a U-shaped kink in the middle for coupling to the tuning coil. Super-regeneration is obtained by increasing the grid-leak resistance until the characteristic rushing sound is heard. Such a receiver, with an anode milliammeter, can be used as a sensitive

* Such as is described by Essen and Gordon-Smith in " The Measurement of Frequencies in the Range 100 Mc/s to 10,000 Mc/s ". *J.I.E.E.*, Part III, Dec. 1945.

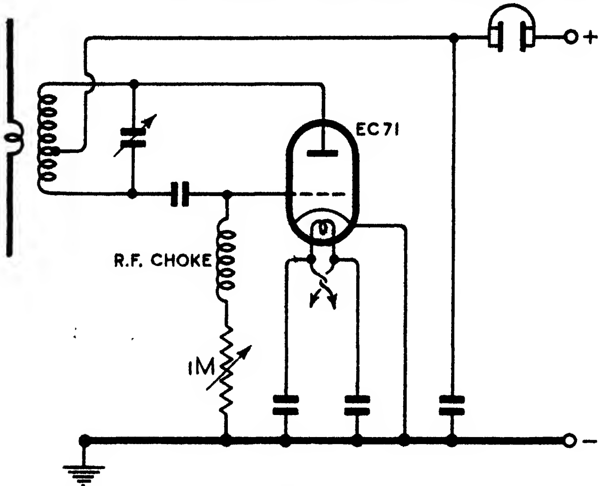


Fig. 12.7—Typical super-regenerative circuit, in which the variable resistor is adjusted to obtain interruption at supersonic frequency of the v.h.f. oscillation

indicator for comparing an unknown signal with a standard signal generator.

12.9. INDICATORS

Many of the commercially available valve voltmeters are designed to cover part or even the whole of the v.h.f. band, by virtue of the carefully designed probe bringing the diode rectifier right up to the "work" (Sec. 5.13). Miniature diodes with small clearances, such as the 6D1, are suitable. The maximum useful frequency depends chiefly on two sources of error. One is series resonance of the head—inductance of connecting spike with capacitance of diode. As the resonant frequency is approached, the voltage actually at the diode becomes greater than that at the point of contact, causing the meter to read high. The actual resonant frequency of a well-designed head may be as much as 1,000 Mc/s, but the error becomes excessive at a few hundred Mc/s. The other error is transit time, which causes the meter to read low, and therefore tends to offset the first, but unfortunately it depends very largely on voltage, as shown in Fig. 12.8. See also the summary of diode-voltmeter errors in general referred to in Sec. 5.13. Even under conditions where the actual voltage is uncertain, a valve voltmeter can be very useful for those measurements in which only comparison is necessary.

Crystal (semiconductor) diodes of suitable types can be used at higher frequencies than the thermionic kind, need no cathode heating (and consequently are free from risk of bringing in hum), have lower forward resistance, are smaller and more robust, and pass no current without input. On the other hand, the types suitable for v.h.f. have comparatively low backward resistance, low maximum backward voltage, and their characteristics are not sufficiently stable for use in measuring instruments without periodical check. Their lower input impedance matters less at v.h.f., because impedances generally tend to be low.

Junction diodes are unsuitable because of their capacitance. In point-contact types, 1 pF is a typical figure.

For the highest frequencies, up to 10,000 Mc/s and even beyond, great numbers of silicon diodes have been produced as frequency changers in radar and other microwave equipment. They are put up in coaxial capsule form, which may be inconvenient at the lower (v.h.f.) frequencies (if one has in mind the use of Government surplus diodes) but not an insuperable difficulty. Another point is that they are easily and quickly ruined by excessive forward current or backward

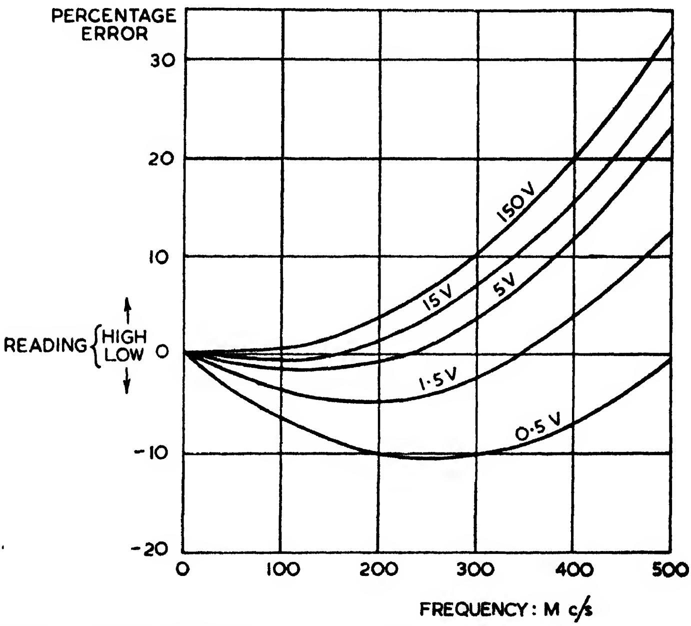


Fig. 12.8—Valve-voltmeter error at v.h.f., due to combined effect of transit time and input resonance, at five indicated r.m.s. voltages. (General Radio type 1800-A.)

voltage. With their indifferent stability, use for comparison of signals is preferable to measurement by once-for-all calibration.

Some, but by no means all, point-contact germanium diodes are suitable for v.h.f. working. They have a lower voltage drop than silicon, and are usually produced with wire connections, which must be soldered with care, using pliers as a heat shunt between the soldering iron and the diode.

For v.h.f. purposes, use has been made of the supremely simple arrangement shown in Fig. 12.9, in which the rectifier serves as its own load resistance and the capacitance of the screened lead is the capacitor, which may be augmented for lower frequencies as shown dotted. Obviously this method of use is unsuitable where there is a d.c. component in the input, or where it has no d.c. path. If the dotted capacitor is used, the length of the screened lead must not approach an odd number of quarter-wavelengths, as it would then present a high impedance at the diode.

The cathode-ray tube can be used up to very high frequencies, but it may be necessary to use rather more anode voltage. The reason is

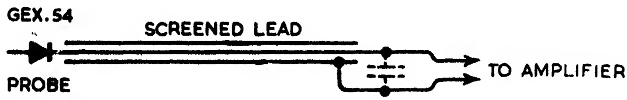


Fig. 12.9—Simple germanium rectifier head for v.h.f. voltmeter

that if the time taken by the beam electrons to pass between the deflection plates is an appreciable fraction of one cycle of the deflecting voltage, there will be a loss of deflection sensitivity (Sec. 14.29). This loss reaches its maximum when the time spent under the influence of a pair of plates is equal to one whole cycle (or any whole number of cycles), for then the deflection due to one half-cycle is neutralized by the opposite deflection due to the other and the sensitivity is nil. And the time taken in passing from Y to X plates introduces a phase shift that might be misleading if the cause were not known.

Travelling-wave c.r. tubes are made (by 20th Century Electronics) for use up to about 1,000 Mc/s and even higher.

12.10. IMPEDANCE MEASUREMENT

The frequency limit of impedance bridges has been pushed so far as to include most of the v.h.f. band. Notable examples are those developed by the B.B.C. and manufactured by Wayne Kerr. They are based on transformer ratio arms (Sec. 7.10). What at lower frequencies is a detail becomes one of the main preoccupations—the design of the terminals so as to define clearly the points between which the impedance is measured. The General Radio Type 1606-A r.f. bridge, though not of the transformer ratio arm type, relies on a special wide-band generator-to-bridge transformer to cover the frequency range 0·4 to 60 Mc/s (*G.R.E.*, June 1955). The Marconi Instruments Type TF978 v.h.f. admittance bridge is a capacitance ratio type covering 30 to 300 Mc/s, in which a thermistor is used as a conductance transfer standard. This bridge and its accessories are described in *M.I.*, Dec. 1959.

A somewhat different system, not commercially available, which was described by J. E. Houldin on behalf of the G.E.C. Research Laboratories in *Proc. I.E.E.*, Part III, Nov. 1952, covers the range 1 Ω to 100 kΩ at 50 to 500 Mc/s by comparison with a standard 100-Ω resistor.

Measurements at frequencies at which lumped-circuit technique

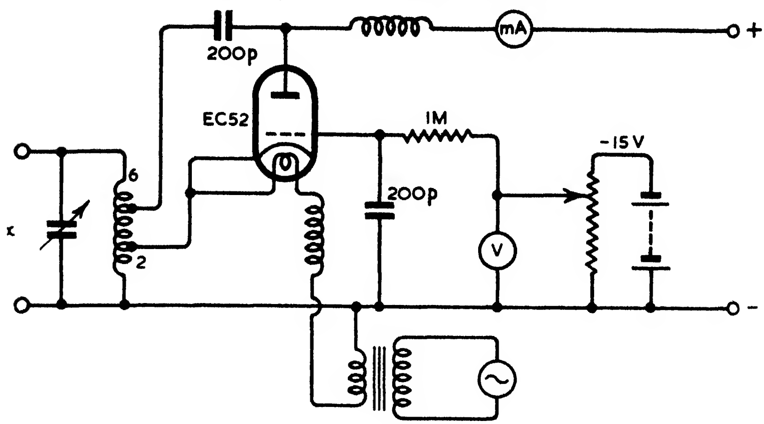


Fig. 12.10—Circuit diagram of triode substitute for dynatron in impedance measurements at v.h.f.

ceases to be practicable—about 150 Mc/s—are discussed in Hartshorn's *R.F. Measurements*, Ch. XII.

At the lower v.h.f., however, some of the methods described in Secs. 9.25 to 9.34 can be used, if sufficient care is taken with circuit layout and components. The Hartshorn and Ward apparatus (Sec. 9.28), for example, is effective up to 100 Mc/s. Generally speaking, miniature components should be used, and leads eliminated from r.f. circuits as much as possible. The resonance-curve methods of measuring Q and r.f. resistance, especially that described at the end of Sec. 9.28, avoid some of the difficulties.

The frequency limit of the dynatron is discussed by G. A. Hay (*W.E.*, Nov. 1946). He shows that it does not fail as an impedance-measuring device because of transit time below 100 Mc/s, but that care is needed in minimizing the effects of external leads. When this has been done, the limit of accurate use is determined by internal reactances; with ordinary valves, around 75 Mc/s. There is also difficulty in bringing any but low-loss circuits to the point of oscillation. If the dynatron fails on this account, the alternatives in Secs. 4.21 and 4.22 may be tried; or, sacrificing two-terminal connection, most of the measurements for which the dynatron has been specified can be performed with a triode oscillator.

Fig. 12.10 is the circuit of an oscillator that has been used successfully for substitution measurements in Band I. The coil consists of six turns of 16 s.w.g. bare copper, wound as a self-supporting solenoid $\frac{1}{2}$ in. in diameter and tapped at two and five turns. The grid bias at which oscillation starts—indicated by a sudden change in anode current—is read on a voltmeter having a large open scale. The anode voltage must be kept very constant.

For measurements on v.h.f. cables, see Sec. 11.2.

CHAPTER 13

Dealing With Results

13.1. REARRANGING FORMULAE

THE ultimate value of laboratory work on the bench often depends largely on what is done with it at the desk. Instrument readings may say too much or not enough if they are improperly or inadequately handled. And it is remarkable how apparently scanty data can sometimes be made to yield information of great value if put through a scientific " third degree " examination.

One of the considerations when devising a method of measurement should be, if possible, the direct presentation of the results; but notwithstanding this it is often necessary to derive them by computation, and sometimes this may be very laborious. The more results there are to be worked out, the more trouble is justified in prearranging the formula.

Take a simple example. The equation connecting the inductance, capacitance and resonant frequency of a tuned circuit can be written in many ways. Of these,

$$\omega^2 LC = 1$$

is the tidiest-looking. But it is the least well adapted for practical purposes. Suppose one is working out the capacitances in a circuit of known inductance that is found to resonate at certain frequencies. Suppose, too, that the inductance happens to be 16 μH, that the frequency is measured in Mc/s, and it is desired to have the capacitance in pF. Unless the contrary is stated, it is assumed that formulae are in basic units. So far as any branch of electrical engineering is concerned —and that includes us—basic units are, by international agreement, the m.k.s. system of units (Sec. 14.1). These include what used to be called the practical electrical units—ampere, henry, farad, etc. For working out results, however, they are sometimes very far from practical! Who would choose to substitute 0·000016 H and 7,500,000 c/s in a formula to get 0·000000000028 F?

Taking first L, which stands for the number of henries, we shall obviously find it more convenient to work in a smaller unit. We could devise a symbol, say $L_{\mu H}$, to stand for the number of microhenries. As there are one million (10^6) microhenries in a henry, $L_{\mu H} \equiv 10^6 L$, so $L \equiv 10^{-6} L_{\mu H}$. ($\equiv$ means " identically equal to";

that is to say, the equality does not depend on circumstances). Therefore we do not affect the truth of the original formula by writing $10^{-6} L_{\mu H}$ in place of L. Similarly with any other factors whose basic units are not convenient. But it would be almost as inconvenient to have to use symbols like $L_{\mu H}$ every time, so provided we remember (or better still make a note of) the fact that L is in microhenries we can use it unadorned. Inserting the appropriate conversion factors in the original equation, thus:

$$(2\pi f \times 10^6)^2 \times 10^{-6}L \times 10^{-12}C = 1$$

and rearranging in terms of C, we get

$$C = \frac{10^6}{4\pi^2 f^2 L} = \frac{25{,}330}{f^2 L} \qquad [\text{pF; } \mu\text{H; Mc/s}]$$

This form is suitable if there are various values of inductance, but if many readings are taken for one value of L, say 16 μH, the work is facilitated still more by reducing the equation to

$$C = \frac{1{,}583}{f^2} \qquad [\text{pF; Mc/s; } L = 16\,\mu\text{H}]$$

When formulae are adapted for special units this should be made clear by a note to that effect, as illustrated above. There is likely to be confusion if the limitations of a formula are not made clear.

13.2. WHAT MAY BE NEGLECTED

Often a formula can be greatly simplified by neglecting something that is too small to be serious within the scope of the work. Suppose the capacitance C of a circuit that oscillates at a frequency f is increased by the small amount ΔC, causing a corresponding small reduction Δf in frequency. Then

$$\Delta C = \frac{C\Delta f(2f-\Delta f)}{(f-\Delta f)^2}$$

This can be used as a method for measuring small capacitance changes (Sec. 9.34). The formula is a most awkward one to calculate, however, for subtractions cannot be done on a slide-rule but have to be worked out beforehand. But if Δf is very small compared with f it can be neglected in terms containing both, so

$$\Delta C \simeq 2C\Delta f/f \qquad [\Delta f << f]$$

—a delightfully easy piece of work.

Very often these approximate formulae are written with the sign of equality (=), and no doubt it is sometimes pedantic to use the sign of approximate equality as above, but it is another of the little things that help to prevent mistakes; and so is the note seen on the right, signifying that to justify the approximation Δf must be much less than f.

The above example illustrates another point, that elimination of "negligible" terms must not be done indiscriminately. Because Δf

is small it is not to be taken as ground for simplifying the equation still further to

$$\Delta C \simeq 0/f$$

It is only as part of a much larger quantity that it can be neglected. Before dismissing a quantity as negligible one must always ask oneself " Negligible in comparison with what? "

13.3. DECEPTIVE FORMULAE

There is a rather more subtle pitfall. It can be illustrated by a somewhat similar example. Suppose one possesses an accurately calibrated capacitor, but no suitable standard inductor. As there are inevitably certain unknown capacitances (such as that of the coil) in shunt with the standard capacitor, the total circuit capacitance is unknown in spite of the calibration, and one is not much farther forward. But if the capacitor readings, C_1 and C_2, required to tune to two known frequencies, f_1 and f_2 respectively, are noted, the extra circuit capacitance, C_0, can be calculated without a knowledge of the inductance:

$$C_0 = \frac{C_1 f_1^2 - C_2 f_2^2}{f_2^2 - f_1^2}$$

This looks quite harmless, and is perfectly correct, but it is a type of formula to beware of. It involves the differences between relatively far larger quantities. Consequently, unless these quantities are known and calculated with extreme accuracy, the answer is not accurately given. The small error due to slide-rule calculation of the large quantities might exceed the final answer! In other formulae there is sometimes a temptation to eliminate a " negligible " term, only to find that, because the formula involves relatively small differences between larger quantities, the despised term was vital after all.

Subject to these precautions, it is usually worth while doing something to adapt a formula to the particular work in hand. It is a good thing to keep a notebook of pet formulae that have proved their value. Their origins and limitations should always be clearly indicated.

13.4. AIDS TO CALCULATION

It is not always necessary to work out results from formulae; many can be derived quickly with moderate accuracy from nomograms (also called alignment diagrams or abacs). A useful collection of these charts for radio work is available as a book.* By laying a straight-edge or stretched thread across the diagram to intersect scales of the known quantities at the appropriate points, the desired quantity can be read off another scale.

The two most generally useful abacs of this type are given here as

* *Radio Data Charts*, by R. T. Beatty and J. McG. Sowerby (Iliffe).

Figs. 14.2 and 14.9. The first, connecting volts, milliamps, ohms, mhos and watts, enables the remaining quantities to be read off when any two are known. Besides covering applications of Ohm's law, it shows whether a proposed resistor is likely to be overheated, the least resistance that is safe for a resistor of given wattage rating and voltage drop, and many other everyday problems. The other diagram connects reactance, inductance, capacitance, frequency and wavelength. Among its many uses are the indicating of reactances, resonance frequencies, required tuning components, and amplifier frequency bands. Even when not accurate enough for the work in hand, the charts are valuable for rough checking of calculations.

The ordinary graph is another form of diagram that can be very useful as an aid to calculation. Ordinarily a single curve serves to relate only two quantities; a " family " of curves is needed for three, and then one of the quantities is represented only by a series of isolated values. Sometimes a single curve can be made more general by scaling the ordinates in quantities that include two or more factors. Some examples of this are given in Chapter 14.

Graphs are almost the only practical way of dealing with problems involving quantities, such as valve characteristics, that cannot be expressed algebraically. There is a whole book on graphical valve calculation.* But even algebraic work is sometimes more easily done —and certainly more easily visualized—graphically, especially if use is made of special kinds of graph paper such as logarithmic paper (Sec. 13.10).

Whatever is available in the form of quick-reckoning diagrams, there are always plenty of things to be worked out; and for this purpose no engineer could possibly go through life without a slide-rule. It was with slide-rule working in mind that the first formula given for ΔC a few paragraphs earlier was viewed with distaste. The simplified formula, on the other hand, can be worked out with one setting. There are a number of publications giving instructions on the multitudinous uses of the slide-rule. For radio purposes one provided with a uniformly divided scale (usually marked " L " or " LOG ") is very helpful for converting ratios to decibels and vice versa (see Sec. 14.34), especially if this scale is on the front.

A " log-log " scale is useful for dealing with formulae embodying logarithms.

The commonly-occurring vector type of calculation ($\sqrt{(X^2+R^2)}$) unless modified as explained in Sec. 14.12, is awkward on the slide-rule.

The accuracy (assuming adequate workmanship) depends on the length of scale, and there are various devices for getting a long scale in compact form. The ordinary straight rule still retains its popularity, but a circular rule has certain advantages, one being that the scales are continuous.

* *Graphical Construction of Vacuum Tube Circuits*, by A. Preisman (McGraw-Hill).

13.5. FALSE ACCURACY

Even quite experienced students or technical assistants are sometimes guilty of recording readings of ordinary meters (full-scale, say 100) such as 86·75 and using them as if the instrument really were accurate to 0·01 % instead of perhaps 1 %. Or, having performed a measurement of r.f. resistance by a method that can be depended upon within perhaps 5 %, they state the result as 34·22 Ω.

These are examples of false accuracy. The manner in which a value is presented ought to indicate its probable accuracy. For example, although "1·4" and "1·400" are numerically equal, they convey a different meaning. The first means anything from 1·35 to 1·45; the second, being much more precise, implies correspondingly greater accuracy, between the narrower limits 1·3995 and 1·4005. Similarly a value stated as "572,000" implies an inaccuracy of only $\pm 0{\cdot}5$. If in fact only three figures are reliable, it should be written as "$5{\cdot}72 \times 10^5$". An alternative method that has been used by the N.P.L. is to drop to a slightly lower level the first figure that is not known with certainty, as for example "$10{,}2_{50}$", which indicates that the greatest possible error to which this result is subject would not affect the first three figures but might affect the fourth. When it is necessary to give a more definite indication of the probable error it should be written, for example, "10,250 ± 25". In assessing the error it should be remembered that relative values are often more accurate than absolute values, so it may be justifiable to take down readings more precisely than their absolute accuracy warrants, in case constant errors cancel out in the final result.

If several factors are multiplied in the ordinary long arithmetical manner, the answer appears with a great many more figures than any of the factors. To the unthinking, it might appear that the answer is known to more decimal places than any of the data used to find it. This, of course, is absurd; and to avoid implying greater accuracy than is justified the surplus figures should be cut off or replaced by noughts, or the shortened methods of multiplication that waste no time finding these meaningless figures should be used. That is not to say that during the calculation it is never necessary to work to more places than will be justified in the answer; an example to the contrary was given in Sec. 13.3. The important thing is to discard misleading digits in the final answer.

In rounding off final figures, there is no doubt that the three-figure value of, say, 2·648 is 2·65; but what about 2·645? The rule with 5 is to round off to an even number, which in this case would be 2·64.*

The slide-rule automatically gives a uniform degree of precision. If it can be read accurately to three figures, then the answer is given with that number, no matter how many factors compose it. For most engineering purposes, the precision of properly effected slide-rule calculations is rather better than the accuracy of the data, or the

* B.S. 1957 : 1953, *The Presentation of Numerical Values* (*B.S.I.*).

accuracy with which the answer need be known. Where this is not so, and data of high accuracy are available for giving an answer that must be known to the same order of accuracy, either the calculation must be done laboriously by arithmetic, or tables of the appropriate number of figures must be used. Incidentally, a book of seven-figure tables is quicker to use to five figures than a book of five-figure tables.

A very large choice of mathematical tables exists—there is, in fact, a comprehensive index of them, by Fletcher, Miller and Rosenhead (Scientific Computing Service, Ltd., 23, Bedford Square, London, W.C.1). Perhaps the best cheap four-figure set is *Chambers's Four-Figure Mathematical Tables*, by L. J. Comrie (Chambers). An exceptionally comprehensive set, with the advantage of having all six trigonometrical functions side-by-side, is *Standard Four-Figure Mathematical Tables*, by L. M. Milne-Thomson and L. J. Comrie (Macmillan). An aid to vector calculations is *Tables for Converting Rectangular to Polar Co-Ordinates*, by J. C. P. Miller (Scientific Computing Service, as above).

13.6. ELIMINATING ERRORS

There is inevitably some error due to the instruments used for making a measurement, and mathematical methods have been devised for combining data in such a way as to indicate the most probably accurate result. For these, a book dealing with the theory of errors should be consulted. Some instrumental error, as has just been said, is inevitable. But there is no excuse for increasing it by faulty working out. One cannot take too much care about this. It is at least an immense waste of time to spoil good work by mistakes in recording or computing. And if one intends to earn a living by technical work it is important to know that in the long run reliability is more valuable than careless brilliance. To this end, columns of figures should be inspected for obvious inconsistencies, rough checks should be made of calculations to ensure that the decimal point is right, and cross-checks devised to arrive at the same results by other routes.

Ref: *Electrical Measurements and the Calculation of the Errors Involved*, by D. Karo (Macdonald, 1950).

13.7. TABULAR WORKING

In general, the soundest method of recording and working out results is in tabular form. By taking the calculation a step at a time the chance of mistakes is reduced, tracing miscalculations is much easier and quicker, and the intermediate steps may show up useful and perhaps unexpected relationships. The alternative of writing down only the final answers, or of working them out on odd scraps of paper that are lost or destroyed, is most exasperating when it is found or suspected that something has gone wrong. Moreover, when work is being referred to by somebody else, or at a much later date, it can be accepted more confidently if the details of how the results have been arrived at are clearly shown.

As an example of tabulation, suppose that a number of measurements have been made, by the method described in Sec. 9.22, of the effect of z.f. current on the inductance and resistance of an iron-cored coil. The formulae are:

$$L = \frac{2T^2M}{r^2+4\omega^2M^2} \qquad R_L = \frac{T^2r}{r^2+4\omega^2M^2} - T$$

where $\omega = 2\pi f$ as usual. Suppose $T = 1{,}000$, $\omega = 5{,}000$, and the readings, M, of the inductances are in μH. Then first the formulae are simplified on these assumptions to

$$L = \frac{2M}{r^2 + \left(\frac{M}{100}\right)^2} \text{ henries and } R_L = \frac{10^6 r}{r^2 + \left(\frac{M}{100}\right)^2} - 1{,}000 \text{ ohms.}$$

The working-out of the readings is then arranged thus:

READINGS I (mA)	M (μH)	r (Ω)	r^2	$\left(\frac{M}{100}\right)^2$	$r^2+\left(\frac{M}{100}\right)^2 = A$	$\frac{2M}{A} = L$(H)	$\frac{10^6 r}{A} = R_L+1{,}000$	R_L (Ω)
0	4,740	2·8	7·8	2,245	2,253	4·2	1,242	242
10	5,120	3·2	10·2	2,630	2,640	3·9	1,210	210
20	5,680	3·9	15·2	3,220	3,235	3·5	1,205	205
30	6,400	4·9	24	4,100	4,124	3·1	1,188	188
40	8,550	8·6	74	7,300	7,374	2·3	1,180	180
50	12,200	17·2	296	14,900	15,196	1·6	1,132	132

The next thing is to draw a graph of L and R_L against z.f. current, I (Fig. 13.1). The convention is to make the independent variable (I in this case) the abscissa. Any points shown on a graph should represent experimental results, not purely mathematical values.

The points referring to the a.c. resistance of the coil are rather irregular, and it will have been noticed in the working-out that a small error in data or calculation causes a relatively large error in the derived value of R_L. If it is important to know with accuracy the variation of R_L with z.f. current it will be necessary to repeat the observations of r and the calculations more carefully.

In the above tabulation use has been made of an abbreviation "A" for one part of the formula. This method is especially helpful in a complicated formula where the same "sub-assembly" occurs more than once, or where it has to be repeated frequently in some working.

13.8. INTERPRETATION OF RESULTS

The whole of any work coming within the scope of this book is in vain if it does not enable some conclusion to be drawn. Generally a measurement is not an end in itself, but is made with the object of deducing something from it about the object tested, and has to be *interpreted*.

Looking at the example just given, it seems clear that the inductance falls off considerably as the z.f. current is increased. As it happens, that is such a well-authenticated phenomenon that it causes no

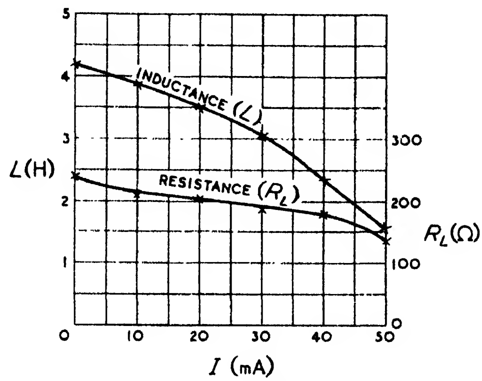


Fig. 13.1—Example of experimental results plotted to aid interpretation

surprise and there is no reason to doubt the validity of the conclusion. But in general one should be extremely cautious in coming even to such a simple conclusion as this. What about the 800-c/s signal current? Was its strength measured, or arrangements made to keep it constant or governed by some simple relationship? If the inductance depended very critically on the a.c., an indefiniteness in this factor might lead to very wrong conclusions quantitatively, and perhaps even qualitatively, about the relationship between inductance and z.f. current. Has the experiment been carried out in such a way as to introduce no other factor into the results? What about the resistance of the battery used for supplying the z.f. current, and the impedance of the milliammeter used for measuring it? Was the frequency of the a.c. *quite* steady throughout? (The results depend on the *square* of the frequency, so this is important.) Such questions as these have to be considered and answered satisfactorily before conclusions can safely be drawn. Suspect everything. The genuine experimenter has a permanently suspicious mind.

He has already suspected that the curious wavelike distribution of the R_L readings may not represent a physical phenomenon but merely insufficient accuracy; and on this assumption a smooth curve has been drawn through them. If the experiment is repeated more accurately and the curve found to droop in a manner similar to that for inductance, the interpreter at once looks for a connection between the two. Obviously the coil's impedance must also droop in the same way. The ratio of resistance to impedance is the power factor: is *that* constant? If so, it is a more interesting result than just to know that inductance and resistance decline in the manner shown by a pair of curves. If it could be established for coils in general (which, as it happens, it cannot) it would mean that if the resistance were known for one inductance and frequency it would be known for all inductances at that frequency. So the next thing is to draw a curve of power factor against z.f. current, to see if it is constant. If not, and no obvious relationship can be perceived, perhaps a curve of power factor against inductance might show one? And so on.

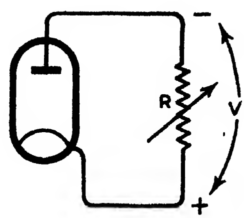
13.9. LAWS

What has just been described is an attempt to derive from some experimental results a general law. This is one of the most important and valuable objects of experiments, so it will be as well to realize what it means. Science assumes that everything in nature works consistently. That is to say, anything that ever happens always happens, given the same conditions. Whenever a current of 2 A is passed through 100 Ω, heat is given off, and always the same amount of heat per second. Furthermore, different happenings are not all completely independent; they are related by fixed principles or *laws*. The law of electrical heating having been discovered, it is no longer necessary to do a special experiment to find how much heat would be given off if 4 A were passed through 50 Ω; it is known that it would be exactly twice as much as with 2 A through 100 Ω. Obviously a knowledge of natural laws saves an enormous amount of time and expense. Without it, every detail of every new engineering work would have to be arrived at by laborious trial and error.

So there is no doubt about the value of discovering such laws. But in attempting to do so one must beware of jumping to unjustifiable conclusions. Do not enunciate a law that All Negroes Have Curly Hair, when the only basis for such a statement is that all the negroes one can remember seeing had curly hair. Reasoning from the particular to the general in this way is legitimate only when the cause of the curliness has been discovered. If careful study of the negroes at one's disposal showed conclusively that the condition of being a negro necessarily caused the hair to curl, then it could be regarded as a natural or scientific law, true of all the millions of negroes one had not seen.

Establishing a scientific law is usually enough to make one's reputation, and few of us can hope to achieve even a minor success at it. The difficulty is making quite sure of identifying the necessary and sufficient cause of the observed effect. That is not to say that nothing worth while can be done short of conclusive proof. If one had observed large numbers of negroes without ever seeing a single exception, the coincidence might be considered sufficiently striking to be put forward as a sort of provisional law—called a hypothesis or an empirical law. A hypothesis is some theory which might eventually turn out to be a natural law. But its provisional status would have to be remembered so long as one could not tell why the two things went together. Coming across a single negro with naturally straight hair would overthrow the hypothesis. The term "empirical law" often means something that is not really a law at all, or even a hypothesis, but just an approximate generalization of experimental results or observations. According to what is called Bode's law the distances of the planets from the sun are approximately proportional to the sequence of numbers obtained by adding 4 to 3 times 0, 1, 2, 2^2, 2^3, etc. Nobody suggests that this has anything to do with a law of nature, or is more than an interesting coincidence, helpful perhaps in aiding the memory. Coming to something more in our

Fig. 13.2—If the electrodes of a diode are joined through a resistance R a voltage V appears across it. The relationship between R and V in a typical sample is plotted in Fig. 13.3



line: if the anode of a thermionic diode is connected through a variable resistance R to the cathode, as in Fig. 13.2, a voltage V appears across it; and if V is plotted against log R the graph is found to be almost exactly a straight line over a wide range of R—say 0·1 to 100 MΩ. This is useful to know when designing valve voltmeters, even though it may be only an empirical law. (There is reason to believe it is more than that.)

13.10. ESTABLISHING LAWS

Plotting graphs is one of the most effective methods of discovering laws, empirical or natural. A graph of V against R in the experiment just mentioned, plotted on ordinary uniformly-squared paper as in Fig. 13.3, might not suggest any simple connection between the two variables. The object is to find some equation that the observed results fit. If various equations are plotted as graphs, they are found to have characteristic shapes. The equation

$$y = ax + b$$

for example, in which a and b are constants, when plotted for various values of x gives points that all lie on a straight line, the slope of which is equal to a and the starting-point on the y axis is b. The equation $y^2 + x^2 = a$ gives a circle; and so on, as explained in school mathematics. x conventionally stands for the quantity that is deliberately varied in the experiment and y for the resulting effect measured. So if the graph of experimental results is a straight line, such as in Fig. 13.4, the values of a and b in the general linear equation follow, and the particular equation thus arrived at is the law of the

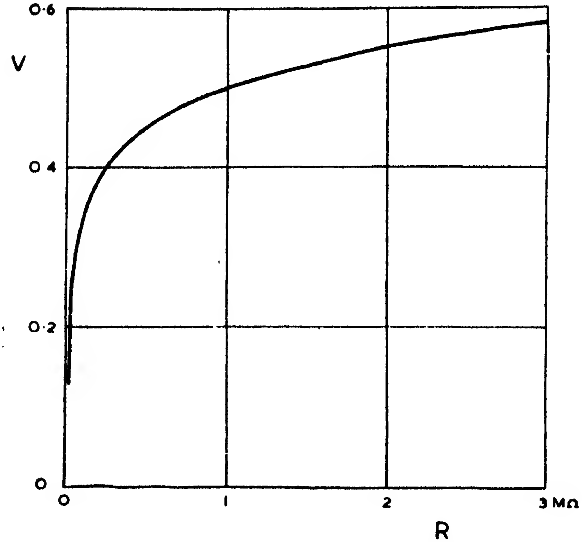


Fig. 13.3—Voltage across a typical diode with normal cathode temperature (Fig. 13.2) plotted against anode-to-cathode resistance, using linear scales

relationship observed, at least over the measured range. In Fig. 13.4, for instance, b is -2, and a, obtained as shown by finding the slope of the line over any convenient section, is 6·5/2 or 3·25. So the particular equation is

$$y = 3{\cdot}25x - 2$$

Whether this is part of some natural law or is only of the empirical kind has to be decided by further investigation.

A straight line is of course by far the easiest shape to recognize and check, so if the " curve " really is curved it is usual to try various dodges such as replotting it on paper ruled logarithmically along one or both axes; if that reduces it to a straight line then its law can be deduced. For example, if the results of the diode experiment are plotted with a log scale for R and a linear (uniform) scale for V, as in Fig. 13.5, an equation can be fitted to it at once. For it is the same thing as plotting log R on a linear scale (which of course is an alternative way of doing it, especially if no suitable log paper is available) and can be solved in the same way as Fig. 13.4, x being log R in this case. The result, very nearly true over the straight-line range, is:

$$V = 0{\cdot}16 \log_{10}R + 0{\cdot}50 \qquad [\text{M}\Omega]$$

The figure 0·50 is found to vary rapidly, but 0·16 only slightly, with cathode temperature.

When some experimental data on the effect of frequency on the input resistance of a certain type of valve were plotted with linear

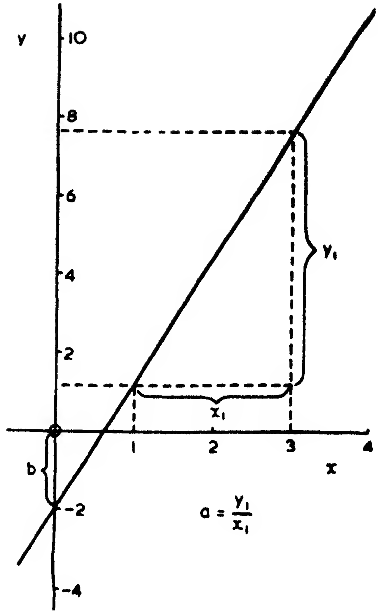


Fig. 13.4—The equation of a straight-line graph is $y = ax + b$; the constants a and b are found as shown in this example

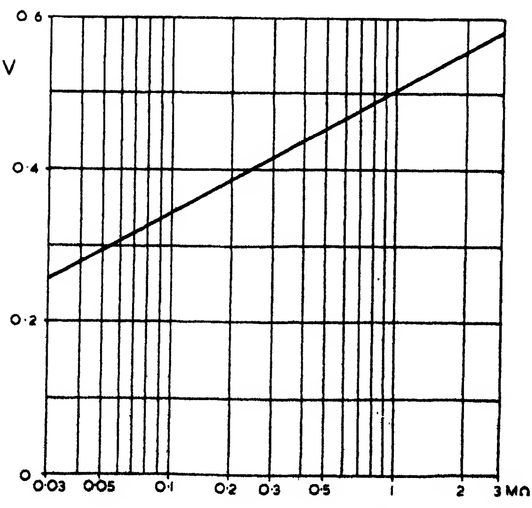


Fig. 13.5—If the curve in Fig. 13.3 is replotted on a logarithmic resistance scale, as in Fig. 13.4, it becomes a very nearly straight line and can be fitted with an equation

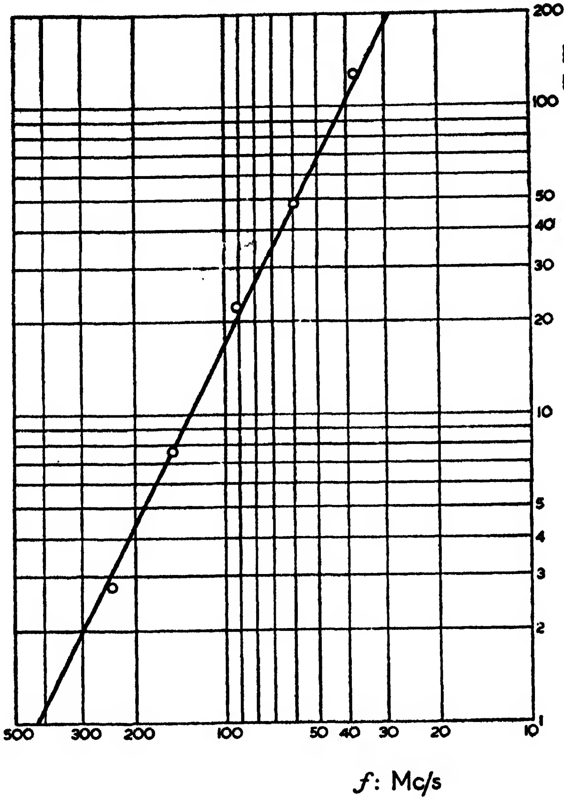


Fig. 13.6—Another example of deriving a law from experimental results ; here the input resistance of a valve is plotted against frequency, both with logarithmic scales; and the slope of the line shows that R is inversely proportional to the square of the frequency

scales the result was an unrecognizable curve. But replotted with log scales, as in Fig. 13.6, the points could reasonably be regarded as lying on a straight line. In a log/log graph, the slope denotes the power or index of x. Here the slope is -2, so it can be deduced that R_i varies as f^{-2}; i.e., is inversely proportional to the square of the frequency. This conclusion can be checked by plotting $1/R_i$ against f^2, which enables the constants to be found. One can say, therefore, that within the limits of accuracy of the experiment, and within the limits of frequency covered, this relationship applies to this particular valve. The constants are likely to be different for different valves or experimental conditions, but if it is found that the main result—$1/R_i \propto f^2$—holds good in many different experiments with many different valves, then one may perhaps venture to work on the hypothesis that it holds for all valves, until evidence to the contrary is forthcoming. Someone meanwhile may have been considering it likely for theoretical reasons that such a relationship should exist. Either empirical law or theoretical proof might come first, and neither alone might be considered quite convincing, but if both agreed (and not because one had been forced or coaxed to fit the other) the case would be secure enough for most people.

13.11. NEED FOR CAUTION

The general warning about not jumping to conclusions can be

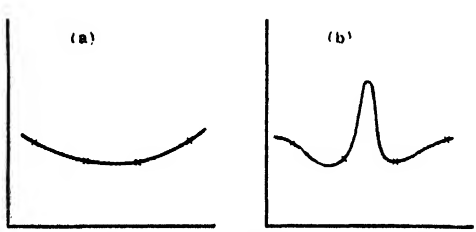


Fig. 13.7—Two ways of drawing a curve through a series of plotted points. Which is right?

Fig. 13.8—If it were not for one point, a straight line could be drawn through them all. Is the odd reading an error, or does it point out a fact?

extended to details such as plotting graphs. Refrain from the temptation to extend a curve beyond the known points (extrapolation), however surely it may seem to tend in a certain direction. The straight line in Fig. 13.5, for example, does not continue the same slope indefinitely towards the left, but tends to flatten out. Even filling in the curve between the known points (interpolation) is not always safe. The curve shown in Fig. 13.7*a* may seem the obvious one to draw through the points given, but closer investigation might show it to be of the form *b*. How about Fig. 13.8? Is the odd point an error in reading, or is it a genuine irregularity in the relationship between the two quantities graphed? The answer must be sought by repeating that particular reading, and preferably also by taking a few extra readings in its neighbourhood. One of the values of a graph is the way it discloses errors, or (alternatively) unexpected phenomena. As F. K. Harris says,* "The mere fact that a result differs by more than one wishes from the others in a set is almost never a sufficient reason for discarding it unless some real explanation can be found for the difference."

13.12. RECORDING RESULTS

Last of all, even if the conclusions or deductions are for immediate consumption it is wise to make a permanent and accessible record of them. This is true even if the answer is, so to speak, a lemon. Negative results are as important as positive, however unsatisfactory they may seem at the time. Results that do not turn out as predicted by theory or previous experiment may show the way to new discovery. If a thing " doesn't work " there is a temptation to throw the job up in disgust; but perseverance, with the object of finding the cause of the failure, is nearly always worth it, if the time can possibly be spared. But however disappointing the result, it is worth recording with full particulars, if only to show at some future date how not to attempt it. There is often a tendency to suppose that work can be remembered and that there is no need to go to the trouble of recording it minutely; or, if it is recorded at all, that rough abbreviated notes will be understood when referred to later. Of course, some people may have entirely different memories, but the author has often looked up notes of work done several years ago and found them either unintelligible altogether or demanding a good deal of study before the threads could be sorted out and the results and the conditions under which they were obtained made clear. Graphs there may be, but without explanations; symbols, with no definitions; and so on.

The best way is to write out an unabbreviated fair copy of the original work; but if that is too much, a few marginal notes explaining what it is all about, and picking out the results prominently so that one has not to wade through a lot of working to make sure which really are the results, is worth while. Unless an experiment is very brief, it must be recorded as it proceeds, and very often some of the earlier parts are later found to be wrong. In such cases a marginal note to that effect should be made, so that when coming to it a long time later the mind does not take on board material only to have to throw it away again when the later results are disclosed.

Another thing: make a careful note of the condition of the experiment—circuits and instruments used. Though some detail among these may seem unimportant at the time, later on it may be realized to have been vital. This part of the work is greatly reduced by adopting the instrument book and serial-number system (Sec. 8.10). Each instrument used can then be referred to by a simple number.

* *Electrical Measurements*, p. 8 (Chapman & Hall, 1952).

Not only is it necessary to record results; one should also make a full note of any ideas that cannot be pursued at the time but that should not be let slip.

13.13. FILING INFORMATION

The actual form of the records depends on individual circumstances. Loose-leaf books, preferably large enough to take graphs without folding, are generally better than plain notebooks, and enable leaflets and data from various sources to be filed in appropriate places. For large quantities of material, foolscap folders in cabinets may be more suitable. There are various more or less elaborate systems of filing by subject, of which by far the most widely used is the Universal Decimal Classification. It has become customary for technical journals to give the U.D.C. numbers of their articles. For one's own filing an elaborate system may tend to become a master instead of a servant, and the entirely unscientific alphabetical arrangement can be more effective for limited purposes. The difficulty about subject filing is that subjects overlap and much mental effort is expended firstly in deciding what subject certain notes should be filed under, and later what part of the system should be searched for those notes.

Information on what other people have been doing is a problem. Much time is certain to be wasted in " discovering " what has already been published, or in preliminary work that would be unnecessary if one were informed about the methods and results of others; on the other hand, it is impossible to read all the world's technical publications, and time is wasted in searching through even a small portion of them. One of the most helpful contributions to the solution of this problem is the Abstracts and References section of *Electronic Technology* (formerly *Wireless Engineer*), and especially the indexes to them published annually.

13.14. COMMUNICATING RESULTS

Sooner or later one's work will have to be communicated to others, in the form of a book, article, paper, report, letter, etc. Some technical workers have supposed—and perhaps some still do—that so long as the information communicated is correct they have done as much as should be expected. The enormous and growing volume of technical information has however made it more than ever necessary that it should be so presented as not to waste reader's time, still less to mislead. This is an art, which calls for extra effort by the writer, but it is better that he make it than that possibly thousands of readers be obliged to do so in order to take the matter in. After all, even authors usually have to spend much more time reading than writing.

Fortunately there is now quite a good selection of books to help one acquire skill in technical writing, and at least one society.* There are

* Presentation of Technical Information Group, University College, London, W.C.1.

also classes at many of the technical colleges. The subject is too big to summarize here, but a few of the commonest faults may be mentioned:

(1) Writers, with the subject matter fully in their minds, often fail to realize that even intelligent readers cannot learn it without being told everything. In particular, they want to know as quickly as possible the purpose, scope and background of the work, so that they can decide whether it is worth their while to read on. Don't plunge into details before making these things clear.

(2) The more inexperienced the writer, the more afraid he is of using a simple, direct, conversational style. He imagines he is under an obligation to be dignified, which means that he tends to be stilted and abstract. So " I made a pinhole in a piece of cardboard and allowed the sunlight to shine through it " (as he might have said in telling someone about it) is transferred to paper as " A parallel beam of light was allowed to pass through a minute aperture in an opaque sheet "—the same number of words, but they give less information and the action is not nearly so easily visualized.

(3) Failure to go through the manuscript again—preferably after a lapse of time, so as to allow the mind to approach it like that of a reader—looking out for ambiguities, misplaced emphasis, unnecessary repetition, omissions, and illogical sequence, non-standard or undefined symbols and units, as well as actual errors.

(4) Causing needless inconvenience and loss of time (and therefore expense) to editor and printer by failing to conform to standard practice in such matters as double spacing, one side of paper only, adequate margins, separate diagrams, standard proof correction practice, etc. This knowledge can be acquired from the books listed below.

Refs:

The Presentation of Technical Information, by R. O. Kapp. (Constable, 1948).

Technical Literature, by G. E. Williams. (Allen & Unwin, 1948).

The Technical Writer, by J. W. Godfrey and G. Parr. (Chapman & Hall, 1959).

Technical Publications, by C. Baker. (Chapman & Hall, 1955).

B.S. 1291 : 1945. *Printers' and Authors' Proof Corrections*. (B.S.I., 1945).

CHAPTER 14

For Reference

14.1. UNITS

FOR a long time students of all branches of electrical engineering were plagued by the fact that there was one system of units—the so-called practical system—for some things, such as e.m.f., current, capacitance and inductance; and two more—varieties of the c.g.s. system—for others, such as m.m.f., magnetic flux and field strength. This mixture has been internationally superseded by the m.k.s. system, based on the metre, kilogram and second, in contrast to the centimetre, gram and second in the c.g.s. system. All the well-known " practical units " —volt, ampere, etc.—fit into this system without change, but new units were necessary to bring into line those quantities that had been reckoned in c.g.s. units.

A disadvantage of the old mixture was that many basic relationships could not be stated without arbitrary constants—powers of 10, and π where it did not correspond with the geometry of the situation. To banish these completely it would have been necessary to alter the size of the ampere and other common units, and make the unit of length the distance travelled by electromagnetic waves through space in one second. As both of these conditions would have been inconvenient, to say the least, the next best thing was done and all illogical constants were incorporated in the values of μ_0 and ϵ_0—the permeability and permittivity of space, or magnetic and electric space constants as they should perhaps be called. Of course they reappear in numerical working wherever μ_0 or ϵ_0 occur, but it is worth while remembering the odd values of these if all equations are otherwise completely clear of illogical constants. (E.g., Sec. 14.8.) The change can be made most simply by regarding the values of μ and ϵ for various materials, based on 1 for empty space, as relative or specific quantities, to be multiplied by μ_0 and ϵ_0 when evaluating equations. (See Table 14.2.)

Disposal of π in this way is called *rationalization*, and was adopted by the International Electrotechnical Commission in 1950, together with the ampere as the fourth basic unit, making the m.k.s.A. system.

In October 1960, the international General Conference of Weights and Measures re-defined the metre as 1,650,763·73 wavelengths of the radiation in a vacuum corresponding to transition between the energy levels $2p_{10}$ and $5d_5$ of the krypton 86 atom. The second was temporarily confirmed as 1/31,556,925·9747 of the tropical year from 0 January 1900 at 12 hours ephemeris time.

This book uses m.k.s. units basically; but where other units such as centimetres are more practical or are still commonly used an alternative formula having the appropriate constants is given, with a note of the units. In the next section the m.k.s. units that differ from those previously used are listed separately for comparison.

Ref: Symposium of Papers on the M.K.S. System of Units, *Proc. I.E.E.*, Part I, Sept. 1950.

14.2. SYMBOLS, ABBREVIATIONS AND UNIT EQUIVALENTS

As far as possible the symbols used in this book are British Standard.* The list that follows includes most of those in common use. To bring out the important distinction between the symbol for a quantity, such as inductance (L), and the abbreviation for the unit in which a quantity is reckoned, such as the henry (H), the former is printed in italics. The latter is correctly used only after a number specifying a particular amount, as 160 μH (standing for 160 microhenries).

Quantities and Units

Table 14.1

Quantity	*Symbol*	*Unit*	*Abbreviation for Unit*
Time	t	second	s or sec
Wavelength	λ	metre	m
Frequency	f	cycle per second	c/s
Electromotive force (e.m.f.)	E	volt	V
Potential difference (p.d.)	V	volt	V
Quantity of electric charge	Q	coulomb	C
Current	I	ampere	A
Power	P	watt	W
Self-inductance	L	henry	H
Mutual inductance	M	henry	H
Capacitance	C	farad	F
Resistance	R	ohm	Ω
Reactance, inductive	X, X_L	ohm	Ω
Reactance, capacitive	X, X_C	ohm	Ω
Impedance	Z	ohm	Ω
Conductance ($= 1/R$)	G	mho	℧ (proposed)
Susceptance ($= 1/X$), inductive	B, B_L	mho	℧ (proposed)
Susceptance ($= 1/X$), capacitive	B, B_C	mho	℧ (proposed)
Admittance ($= 1/Z$)	Y	mho	℧ (proposed)

* B.S. 1991 : Part 1 : 1954. *Letter Symbols, Signs and Abbreviations* (B.S.I.).

The symbols for quantities such as current, voltage, flux density, etc., may be modified as in the following example to indicate the basis of reckoning:

I	r.m.s. value of current
i (or I_t)	instantaneous value
I_{max} (or $\hat{I}$)	peak value
I_{p-p}	peak-to-peak value
I_{av}	mean value
$\mathbf{I}$	vector notation

The following are among the units affected by adoption of the rationalized m.k.s. system:

Table 14.2

Quantity	*Symbol*	*Rationalized M.K.S. Unit*	*Relationship to " C.G.S. cum Practical " Unit*
Mass	m	kilogram	$= 10^3$ gm
Length	l	metre	$= 10^2$ cm
Force	F	newton	$= 10^5$ dynes
Work and energy	W	joule = watt-sec = newton-metre	$= 10^7$ ergs
Electric field strength	$\mathcal{E}$	volt per metre	$= 10^{-2}$ volts per cm
Magnetic field strength	H	amp-turn per metre	$= 4\pi/10^3$ oersteds
Magnetomotive force	F	ampere-turn	$= 4\pi/10$ gilberts
Magnetic flux	Φ	weber	$= 10^8$ maxwells or " lines "
Magnetic flux density	B	weber per sq. metre	$= 10^4$ gauss or " lines "/sq. cm.
Resistivity	ρ	ohm-metre (or ohm per metre cube)	$= 10^2$ ohm-cm
		Rationalized M.K.S. Value	*C.G.S. Value*
Absolute permittivity	ϵ	$\epsilon_0\epsilon_s$	same as ϵ_s
Electric space constant	ϵ_0	$1/(36\pi \times 10^9)$ approx.	1
Relative or specific permittivity	ϵ_s	same as ϵ in c.g.s.	same as ϵ
Absolute permeability	μ	$\mu_0\mu_s$	same as μ_s
Magnetic space constant	μ_0	$4\pi/10^7$	1
Relative or specific permeability	μ_s	same as μ in c.g.s.	same as μ

The following word endings should be noted:

Ending	*Denotes*
—ion	a property or quality
—ance	a quantity of the property
—or	a component providing a desired quantity of the property
—ive	the adjective corresponding to *—ion*
—ivity	the *—ance* possessed by unit quantity of a material.

E.g. *Conduction* is the property of passing current
Conductance is the quantity of conduction
Conductor is a component or part of equipment designed to conduct
Conductive: having conduction
Conductivity is the relative or specific conductance of a material: the conductivity of copper at 20° C is about 58 megamhos per metre cube (*not* per cubic metre!)

Not all names are fully inflected in this way; and what might have been called "capacitivity" is actually "permittivity", and instead of "inductivity" there is "permeability".

Unit Multiple and Submultiple Prefixes

Table 14.3

Abbreviation	*Read as:*	*Multiplies unit by:*
G	giga-	10^9
M	mega-	10^6
k	kilo-	10^3
m	milli-	10^{-3}
μ	micro-	10^{-6}
n	nano-	10^{-9}
p (or μμ)	pico-	10^{-12}

It is not uncommon, in circuit diagrams, for the symbol Ω, F (or even μF), and H to be omitted from the values of components, leaving only the prefixes as above, where applicable. Thus "100" alongside the symbol for a resistor would signify "100 Ω"; "160 μ" alongside a coil, "160 μH"; and so on.

Miscellaneous Units and Equivalents

Table 14.4

Length

1 kilometre (km)	= 0·6214 mile = 1,094 yards = 1,000 metres	1 mile	= 1.609 km = 1,760 yards = 5,280 feet
1 metre (m)	= 1·094 yards = 3·28 feet = 39·37 inches = 100 centimetres = 10^6 microns = 10^{10} angstroms	1 yard (yd) 1 foot (ft) 1 inch (in)	= 0·9144 metre = 3 feet = 36 inches = 30·48 cm = 12 inches = 2·540 cm = 1,000 mils

Table 14.4 (*continued*)

Length				
	1 centimetre (cm)	= 0·3937 inches = 10 millimetres	1 mil	= 0·001 inch = 25·4 microns
	1 millimetre (mm)	= 39·37 mils = 1,000 microns = 10^7 angstroms		
	1 micron (μ)	= 0·001 mm = 10^4 angstrom		
	1 angstrom (Å)	= 10^{-10} metre		
Area				
	1 square metre (m^2)	= 10·76 square feet	1 square foot (sq ft or ft^2)	= 929 cm^2
	1 square centimetre (cm^2)	= 0·155 square inches	1 square inch (sq in or in^2)	= 6·452 cm^2
Volume				
	1 litre (l)	= 61·02 cubic inches = 1,000 cubic centimetres	1 cubic foot (cu ft or ft^3)	= 28·32 litres
	1 cubic centimetre (c.c. or cm^3)	= 0·06102 cubic nches	1 cubic inch (cu in or in^3)	= 16·39 cm^3
Angle				
	1 radian	= $\frac{360}{2\pi}$ or 57·29 degrees	1 revolution	= 2π radians = 360 degrees
			1 degree (°)	= 60 minutes
			1 minute (′)	= 60 seconds(″)
Mass				
	1 kilogram (kg)	= 2·205 pounds = 35·27 ounces = 1,000 grams	1 pound (lb) (avoirdupois)	= 0·4536 kg = 16 ounces
	1 gram (gm)	= 0·0353 ounce	1 ounce (oz)	= 28·35 gm
Force				
	1 newton	= 0·2248 pounds weight* = 10^5 dynes	1 pound weight*	= 4·448 newtons
			1 ounce weight*	= 27,800 dynes
	1 gram weight*	= 981 dynes		
Energy and Work				
	1 kilowatt-hour (kWh)	= 860 kg. calories = $3{\cdot}6 \times 10^6$ joules	1 foot-pound (ft lb)	= 1·356 joules
	1 joule (or watt-second)	= 0·7376 ft lb = 10^7 ergs		
	1 electron-volt (eV)	= $1{\cdot}602 \times 10^{-19}$ joule		
Power				
	1 kilowatt (kW)	= 1·341 horsepower = 737·6 ft lb/sec	1 horsepower (hp)	= 0·7457 kW = 550 ft lb/sec

* *at sea level at 45° latitude*

Temperature

Table 14.5

	Degrees:		
	Kelvin (°K)	Centigrade (°C)	Fahrenheit (°F)
Absolute zero	0	− 273	− 460
Freezing water	273	0	32
Boiling water	373	100	212
At normal pressure			

$°K = °C + 273$ $°F = \frac{9}{5}°C + 32$

Light See Sec. 14.30.

Greek Alphabet

A	α	alpha	H	η	eta	N	ν	nu	T	τ	tau
B	β	beta	Θ	θ	theta	Ξ	ξ	xi	Υ	υ	upsilon
Γ	γ	gamma	I	ι	iota	O	o	omicron	Φ	ϕ	phi
Δ	δ	delta	K	κ	kappa	Π	π	pi	X	χ	chi
E	ϵ	epsilon	Λ	λ	lambda	P	ρ	rho	Ψ	ψ	psi
Z	ζ	zeta	M	μ	mu	Σ	σ or s	sigma	Ω	ω	omega

Mathematical Symbols and Abbreviations (see Sec. 14.33)

$=$ equal to

$\equiv$ identically equal to (i.e., for any values of variables)

$\simeq$ approximately equal to

$<$ less than

$<<$ much less than

$\not<$ not less than

$\leq$ equal to or less than

$>$ greater than

$\propto$ proportional to

∞ infinity

$|\ |$ magnitude or numerical value of term enclosed (i.e., regardless of sign or vectorial direction)

e base of natural logarithms ($=2{\cdot}71828\ldots$).

ln (or $\log_e$) natural logarithm

j complex vector operator (see Sec. 14.12)

$n!$ $1 \times 2 \times 3 \times \ldots \times n$

$\sqrt{\ }$ square root of

$\sqrt[n]{\ }$ n^{th} root of

Note: $10^3=1{,}000$ $10^0=1$ $10^{\frac{1}{2}}=\sqrt{10}$
$10^2=100$ $10^{-1}=0{\cdot}1$ $10^{-\frac{1}{2}}=1/\sqrt{10}$
$10^1=10$ $10^{-2}=0{\cdot}01$ $10^{\frac{3}{2}}=\sqrt{10^3}$
etc.

$f(x)$ or $F(x)$ function of x

$f'(x)$ or $\frac{d}{dx}f(x)$ or $D_x f(x)$

differential coefficient, or slope, of $f(x)$ with respect to x

$\int f(x)dx$ integral of $f(x)$ with respect to x

$\sin^{-1}x$ angle whose sine is x

δ a small difference or increment

Δ a difference or increment, not necessarily small

Π product (of all terms of type specified)

Σ sum (of all terms of type specified)

Graphical Symbols

A full list is not given here; it is available as B.S. 530: *Graphical Symbols for Telecommunications* (B.S.I.).

Most of those commonly used are well known, but a key to the following may be helpful:

indicating instrument (particular type specified by letter enclosed; e.g. V for voltmeter)

microphone

electrolytic capacitor (white plate positive)

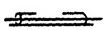
piezo crystal

cell (long stroke positive)

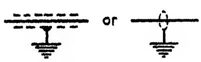
indicator lamp

connection to chassis

dust core



coaxial cable



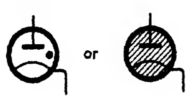
lead with earthed screen

variable

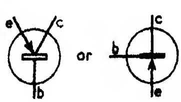
pre-set

adjusted in steps, the number of which may be specified as shown. This last is proposed by the author and is particularly useful for showing range switching of measuring instruments.

thermistor



black spot denotes tube is gas-filled



pnp transistor. *npn* transistor is the same except for arrow reversed

Note: The first symbol is appropriate for point-contact and the second for junction types.

Zener diode

tunnel diode

—	d.c.
$\sim$	a.c. in general
$\eqsim$	d.c. or a.c.
$\backsimeq$	undulating or rectified current
$\approx$	a.f.
$\approxeq$	above a.f. in frequency

Miscellaneous Symbols

A	voltage gain or amplification (alternative to α)
B	feedback factor (alternative to β)
c	speed of light, = 299,792 km/s
e	charge on electron, = $1{\cdot}602 \times 10^{-19}$ coulomb
f_0	frequency of oscillation
f_r	frequency of resonance
Δf	frequency difference
f'	frequency off-tune
g	acceleration due to gravity (at earth's surface), = 9·81 metres/sec^2
h	Planck's constant, = $6{\cdot}625 \times 10^{-34}$ joule-sec
h_e	effective height of aerial
k	coefficient of coupling
k	Boltzmann's constant, = $1{\cdot}380 \times 10^{-23}$ joule/deg.
m	mass of electron at rest, = $9{\cdot}1083 \times 10^{-31}$ kg
Q	circuit storage factor, or ratio of reactance to resistance
R_0	characteristic resistance
Z_0	characteristic impedance
R_x	unknown resistance, to be measured (and similarly for other quantities)
α	attenuation coefficient
β	phase-change coefficient
δ	(1) loss angle, = $90° - \phi$, (2) logarithmic decrement, $\simeq \pi/Q$; (3) a small difference or increment
$\tan\delta$	dissipation factor or loss tangent, = $\cot\phi$, = R/X, = $1/Q$. When $\delta <$ about 10°, $\tan\delta \simeq \sin\delta \simeq \delta$ (in radians)
θ	(1) an angle; (2) temperature; (3) thermal resistance
π	circumference/diameter of circle, = 3·1415926535 . . .
τ	time period
ϕ	phase angle (between vectors representing alternating current and voltage)
$\cos\phi$	power factor, R/Z, = $\sin\delta$

Valve Symbols and Abbreviations

The following are the most important conventions set out fully in B.S. 1409: *Letter Symbols for Electronic Valves* (B.S.I.):

Electrodes

k	cathode
h	heater
f	filament
g	grid
g_1	first grid (nearest cathode)
g_2	second grid; and so on
a	anode
m	internal metal coating
M	external metal coating

Parameters

μ	amplification factor from g_1 to anode
$\mu_{g_1-g_2}$	amplification from g_1 to g_2
r_a	anode a.c. resistance or slope resistance
g_a	anode conductance, = $1/r_a$
g_m	mutual conductance
g_c	conversion conductance (of frequency changer)
R_{eq}	equivalent noise resistance

Symbols are frequently combined, thus:

I_a	anode current
V_{g_2}	voltage at second grid
R_a	resistance connected externally to anode
c_{gk}	internal capacitance from grid to cathode

Capital letters are used for associated items outside the valve; small letters for items inside the valve itself.

When it is necessary to specify a valve interelectrode capacitance in such a way as to indicate precisely how the electrodes are to be connected when measuring it, a system has been standardized in which electrodes to be connected to reference earth (and thereby excluded from the measurement) are mentioned in brackets. Two abbreviations are used:

R remaining elements of the same unit(s), shields, metal parts (e.g., external shields, base sleeves, unused pins or leads)

u inactive units of multiple valves.

For example, "c_{in}" is more precisely denoted by "$c_{g1-R(a)}$", which means the capacitance measured between the control grid and all other electrodes except the anode. And "$c_{gt-R(atu)}$" in a triode-pentode means the capacitance between triode grid and cathode plus heater, the triode anode and all pentode electrodes being earthed.

Transistor Symbols and Abbreviations

A great many transistor parameters have been defined, but unfortunately there is at the time of writing no general agreement either on which to prefer or on their symbols. The most unsatisfactory feature is that whereas all valve parameters, such as μ, are understood to refer to the common-cathode configuration, transistor parameters, such as α, refer to the common-base configuration, so that those for the common-emitter (analogous to common-cathode and, like it, most commonly used) are denoted by such arbitrary modifications as α' or β. The former device (′) is at least of universal application, and ″ can be used to denote common-collector parameters. Until transistors can be brought into line with valves by agreement that unmodified symbols refer to common-emitter, a sound practice is to use a subscript e, b or c, to show which electrode is common to input and output.

In the list below, configuration distinctions are omitted. The only symbol which cannot be used for any of the three, by modification according to one of the alternative systems just mentioned, is the very odd β.

α small-signal current amplification factor

α_0 ditto, at z.f.

$\bar{\alpha}$ large-signal current amplification factor

β current amplification factor with common emitter, $\equiv \alpha' \equiv \alpha_e$

I_{co} Collector current with open-circuited input junction

f_α frequency at which α is 3dB lower than α_0

f_1 " " " the common-emitter α (α', or α_e or β) $= 1$

Typical symbolism for internal elements of transistors is shown in Fig. 14.1—an example of the many equivalent circuits (see also Sec. 14.22).

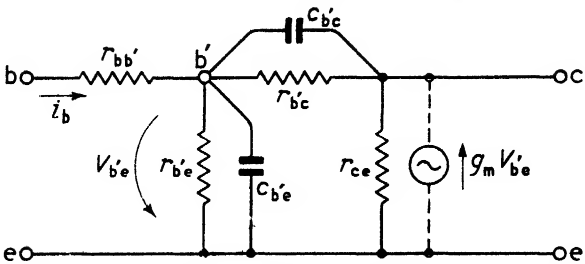


Fig. 14.1—The equivalent lumped internal parameters of a transistor are denoted by symbols as shown here

The following alternative sets of " black-box " small-signal low-frequency parameters are commonly used:

Hybrid Parameters

h_{11} input impedance with output short-circuited to a.c.
h_{21} current amplification factor with output short-circuited to a.c.
h_{22} output admittance with input open-circuited to a.c.
h_{12} voltage feedback ratio with input open-circuited to a.c.

Impedance Parameters

z_{11} input impedance with output open-circuited to a.c.
z_{21} forward transfer impedance with output open-circuited to a.c.
z_{22} output impedance with input open-circuited to a.c.
z_{12} reverse transfer impedance with input open-circuited to a.c.

Admittance Parameters

y_{11}, y_{21}, y_{22}, y_{12}: same as above with " admittance " substituted for " impedance ".

Modified Impedance Parameters

Consist of α, z_{11} and z_{22} with
z_{in} input impedance with output short-circuited to a.c.
z_{out} output impedance with input short-circuited to a.c.

Miscellaneous Abbreviations

a.a.c.	automatic amplitude control
a.c.	alternating current
a.f.	audio frequency
a.f.c.	automatic frequency control
a.g.c.	automatic gain control
a.m.	amplitude modulation
a.v.	alternating voltage
b.f.	beat frequency
c.r.	cathode ray
dB	decibels
dB(mW) or dBm	decibels with reference to 1 mW
d.c.	direct current
d.v.	direct voltage
e.h.t.	extra-high tension
e.m.f.	electromotive force
f.m.	frequency modulation
f.s.d.	full-scale deflection
h.t.	high tension
i.f.	intermediate frequency
l.t.	low tension
m.f.	modulation frequency
m.m.f.	magnetomotive force
p.d.	potential difference
p-p	peak to peak (value)
r.f.	radio frequency
r.m.s.	root-mean-square
s.f.	signal frequency
v.f.	video frequency
z.f.	zero frequency

The use of "l.f." and "h.f." to mean audio frequency and radio frequency is obsolete; it would be liable to be confused with the following nomenclature for *radio* frequencies, agreed at Geneva in 1959:

Table 14.6

Frequency band	*Adjectival abbreviation*	*Frequency range**	*Corresponding wavelengths†*
4	v.l.f.	3–30 kc/s	100–10 km
5	l.f.	30–300 kc/s	10–1 km
6	m.f.	300–3,000 kc/s	1,000–100 m
7	h.f.	3–30 Mc/s	100–10 m
8	v.h.f.	30–300 Mc/s	10–1 m
9	u.h.f.	300–3,000 Mc/s	100–10 cm
10	s.h.f.	3–30 Gc/s	10–1 cm
11	e.h.f.	30–300 Gc/s	10–1 mm

* Lower limit exclusive; upper limit inclusive
† Lower limit inclusive; upper limit exclusive

The above band numbers should not be confused with the roman band numbers given on p. 505.

Although "r.f." includes "i.f.", where both are used together "r.f." is understood to refer particularly to the frequency of the received carrier wave. This is sometimes called "s.f." (signal frequency), but this term has also been used to mean "modulation frequency". Although "v.f." primarily refers to the wide range of frequency with which a television sender is modulated, the term is extended to include analogous frequencies in radar, etc.

Terms that are much misused are "d.c." and "a.c." Where "d.c." is intended to mean "not alternating", and especially where the reference is not really to current, it makes better sense to substitute "z.f." And "d.v." is surely preferable to "d.c. voltage", which is a contradiction in terms and leads to absurdities such as "d.c. current". There is no need, either, to write (as some do) "i.f. frequency"; "i.f." is sufficient.

14.3 Ohm's Law

Although Ohm's law is supposed to be well known, it is often misunderstood. What Ohm said can be expressed as

$$I \propto E \quad \text{or} \quad \frac{E}{I} = \text{constant}$$

and this is not always true (e.g., of thermistors, rectifiers and valves) The familiar expression

$$I = \frac{E}{R} \quad \text{or} \quad E = IR \quad \text{or} \quad R = \frac{E}{I}$$

is true only if a suitable system of units is used (e.g., V, A and Ω, or V, mA and kΩ). It is then universally true, if R is not assumed to be

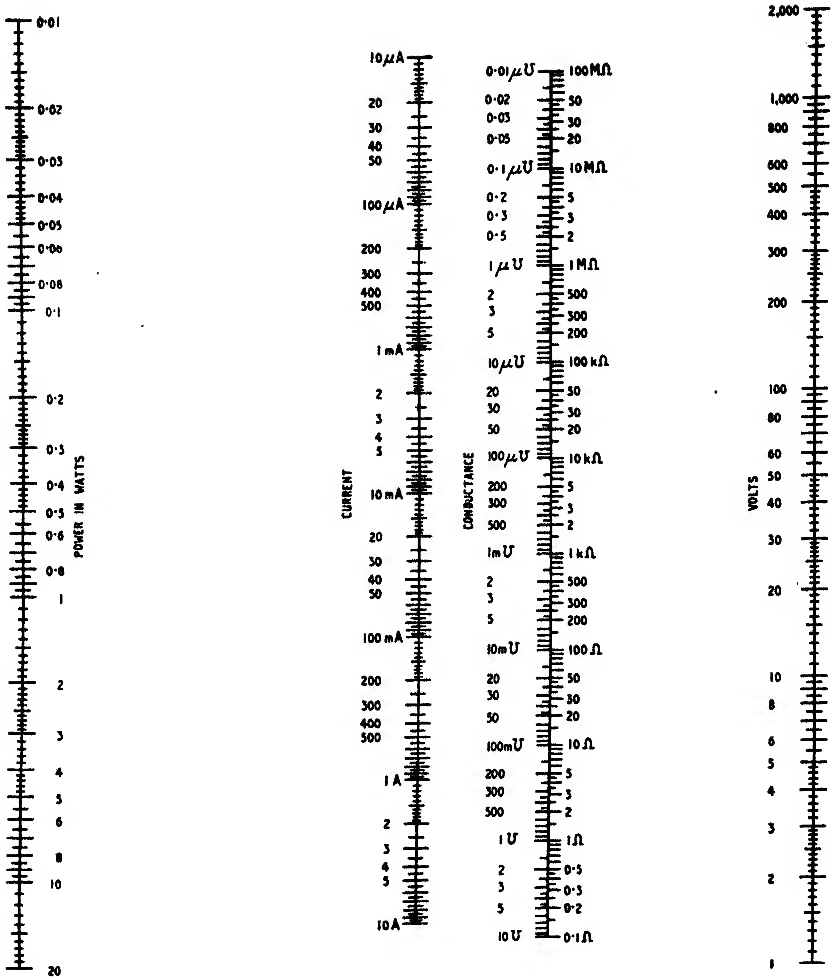


Fig. 14.2—This chart (adapted from No. 42 in Radio Charts, *5th ed.) is used by connecting any two known quantities by a ruler or a stretched thread, whereupon the intersections indicate the other three quantities. E.g., a 10-k*Ω *resistor rated at ½W is seen to be limited to 7 mA and 70 V. Its conductance is also shown as 100μ℧ (℧ being the proposed symbol for " mho "). For an alternative chart, with a preferred value scale of resistance, see* Wireless World, *Jan. 1957, p. 19*

necessarily constant. Since non-linear resistances are now commonplace, and suitable units are universal, it is the latter expression that has come to be called Ohm's law. Though this practice is convenient it is not strictly correct.

An alternative form (Sec. 14.13) is

$$I = EG$$

This is convenient for parallel circuits, for the total conductance, G,

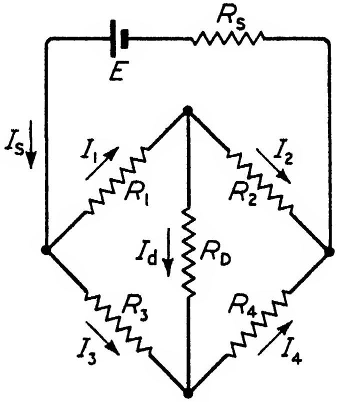


Fig. 14.3—Example of resistance-reactance potential divider

is the sum of the separate conductances in parallel. For relating E, I, R, G and P, see Fig. 14.2.

Ref: " Ohm's Law ", *W. W.*, Aug., 1953.

14.4. KIRCHHOFF'S LAWS

(1) *The sum of the currents flowing to any point in a network is zero.*

If it were not so, a charge of electricity would accumulate at a point, which has no magnitude and so cannot hold a charge. Any currents flowing to the point (positively) must be balanced by negative current (i.e., away from the point).

(2) *The sum of the voltages round any closed path is zero.*

If it were not so, some point would have to be at more than one potential at the same time, which is impossible.

The value of these laws is in solving electrical networks. For example, in the bridge network (Fig. 14.3) one can write down Kirchhoff equations, such as:

$$I_s - I_1 - I_3 = 0$$

$$I_1R_1 + I_dR_d - I_3R_3 = 0$$

The three loops and four junctions give seven simultaneous equations for solving the six unknown currents. Solving them by ordinary algebra is excessively tedious; for such work determinants are used. See also Sec. 14.23.

14.5. RESISTANCE

Resistance of conductor having uniform cross-section area A, length l, and resistivity ρ :

$$R = \frac{l\rho}{A}$$

R, l and A in same units as ρ; e.g., if ρ is in ohm-cm (ohms per cm. cube), l and A must be in cm and sq cm, and R in ohms.

For standard annealed copper, ρ at 20°C is 1·7241 $\mu\Omega$-cm, = 0·0172 $\mu\Omega$-metre, increasing 0·4% per °C.

The comparative resistances of other metals (at z.f.), compiled from various sources, are given in Table 14.7. Some values are liable to differ considerably, depending on the exact composition of the metal.

Table 14.7

Metal	*Comparative resistance*	*Temp. Coefft.*	*Metal*	*Comparative resistance*	*Temp. Coefft.*
Advance	28·4	0·001	Manganin	27	0·001
Aluminium, hard-drawn	1·63	0·39	Mercury	56	0·09
Brass	4·1	0·2	Nichrome	58	0·04
Concordin	60	0·17	Nickel	6·3	0·6
Constantan	28–29	0·0008	Nickel-chromium (80/20)	66	0·0048
Copper, annealed	1	0·4	Nickel-copper (44/56)	32	0·0014
„ , hard-drawn	1·03	0·38	Phosphor bronze	4·4	0·18
Duralumin	2·0	—			
Evanohm	77	0·002	Platinoid	20	0·03
Eureka	28–29	0·0008	Platinum	5·6	0·36
Nickel silver	20	0·04	Silver	0·94	0·38
Gold	1·4	0·34			
Iron, soft	6·1	0·5	Tin	6·7	0·42
Karma	77	0·002	Tungsten	3·3	0·5
Lead	12·8	0·4	Zinc	3·7	0·4

Temperature coefficients are in *percentage* rise in resistance per °C from 20° C.

Thermoelectric e.m.f., copper-Eureka junction: approx. 0·04 mV/°C.

Skin Effect

The maximum gauge of straight wire of which r.f. resistance does not exceed a.f. resistance by more than 1% is shown in Table 14.8.

Table 14.8

Frequency	*Copper*	*Eureka**	*Frequency*	*Copper*	*Eureka**
kc/s	s.w.g.	s.w.g.	Mc/s	s.w.g.	s.w.g.
100	30	15	5	47	33
200	34	18	10	49	37
300	36	19	20	50	40
500	38	21	30	—	42
1,000	42	24	50	—	44
2,000	45	28	100	—	47
3,000	46	30			

* Sufficiently correct for manganin.

14.6. CAPACITANCE

Parallel plates

Basic m.k.s. equation:

$$C = \frac{A\epsilon}{t}$$

C in farads
A in sq metres
t in metres

$\epsilon = \epsilon_0\, \epsilon_s$, where $\epsilon_0 = 10^7/4\pi c^2 = 8{\cdot}854 \times 10^{-12}$ farad/metre.

In other words, a pair of plates, each 1 sq metre spaced 1 metre apart, with uniform parallel field (i.e., no edge effect) has a capacitance of 8·854 pF. With dimensions in centimetres:

$$C = \frac{0{\cdot}0885\, A\, \epsilon_s}{t} \text{ picofarads}$$

If edge effect is neglected, A is the *total* dielectric cross-section, equal in a two-plate capacitor to the plate area; in a three-plate capacitor, to twice the plate area, etc. Values of ϵ_s, the specific permittivity, or dielectric constant, are given in Table 14.9.

Capacitance where more than two conductors are involved—see Sec. 9.16.
Coaxial cylinders, or screened wire—see Sec. 14.27.
Capacitor charge or discharge

If a capacitance of C microfarads, initially charged to V_0 volts, is connected to a source of E volts in series with R megohms, the voltage

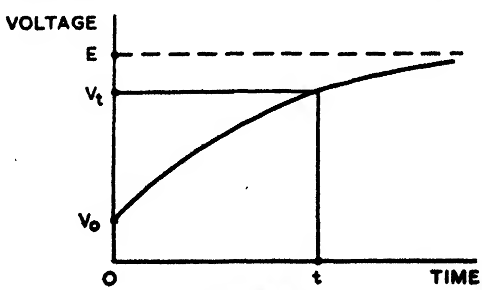


Fig. 14.4

of C after t seconds will be

$$V_t = E - (E - V_0)e^{-t/RC}$$

(e = base of natural logarithms = 2·718 . . .)

Alternatively $$2{\cdot}30 \log_{10} \frac{E - V_t}{E - V_0} = -\frac{t}{RC}$$

When charging from zero, $V_0 = 0$; when discharging to zero, $E = 0$.

The quantity RC is the *time constant*, equal to the time occupied by the first 63·2% (= 1 − 1/e) of the voltage change from V_0 to E.

Energy stored

The energy stored in C farads charged to V volts = $\frac{1}{2} V^2 C$ joules.

14.7. PROPERTIES OF INSULANTS AND DIELECTRICS

ϵ_s = dielectric constant or specific permittivity

tan δ = dissipation factor (nearly the same as power factor; see Sec. 6.2).

ρ = volume resistivity, in *meg*ohms per cm cube.

Table 14.9

Material (*Trade names in brackets*)	ϵ_s	*tan* $\delta \times 10^4$		ρ
			c/s	MΩ-cm
Air, at 0°C and 760 mm	1·0006	practically 0		practically ∞
Casein (Erinoid)	6·2	500	10^6	
Cellulose-acetate film	4–5·5	400	10^6	10^4
Ebonite, unloaded	2·8–2·9	60–80		10^{10}
,, , magnesium-carbonate-filled (Keramot)	3·8–4·1	100	10^6	
Epoxy resins (Araldite)	3–4	150–800		
Glass, plate	6–7	40	10^6	10^7
Methyl methacrylate (Perspex)	2·6–3·4	60	10^6	10^9
		600	50	
Mica	4·5–7	2–3	10^6	10^5–10^{11}
Paraffin wax	2·2	1–2	10^6	10^{12}
Phenol formaldehyde (Bakelite), mouldings	5	300–400	10^6	10^6
Phenol formaldehyde (Bakelite), paper	4·5	300	10^3	10^6
Polyamides (Nylon)	3–4	200		10^7
Polyethylene (Polythene, Alkathene, Telcothene)	2·3	1–3		10^{11}
Polystyrene (Distrene, Trolitul)	2·6	1–5		10^{11}
Polytetrafluorethylene (Fluon)	2·2	2		10^{10}
Polyvinylchloride (PVC)	2·8–5	100–500		10^7
Porcelain	5–7	50	10^6	10^8
Quartz, fused	3·9	2	10^6	10^9
Rutile (Faradex)	80	3	10^6	10^{10}
Shellac	3·5	300	10^6	10^{10}
Silica, fused	3·5	2	10^6	10^8
Steatite (Frequentite, etc.)	6	2–20	10^6	10^{10}
Urea formaldehyde (Beetle)	6	300	10^6	10^6

Some compounds of titanium have ϵ_s up to 3,000 or more.

The values are liable to differ considerably, depending on the exact composition of the material, and its temperature. Where a frequency is not stated for tan δ, the values given cover a wide range of frequency.

14.8. ELECTROMAGNETISM

In m.k.s. units:

$F = IN$	M.m.f. in amp-turns due to I amps flowing through N turns.
$H = \dfrac{IN}{l}$	Magnetic field strength in amp-turns per metre due to IN amp-turns acting over l metres of uniform magnetic path.
$B = \mu H = \dfrac{4\pi}{10^7}\mu_s H$	Flux density in webers per sq metre in medium of absolute permeability μ, or μ_s relative to vacuum as 1.
$\Phi = BA$	Total magnetic flux in webers due to uniform flux density B in magnetic path A sq metres cross-sectional area.
$L = \dfrac{\Phi N}{I} = \dfrac{\mu A N^2}{l}$	Inductance in henries of a coil of N turns in which a current of I amps produces a flux of Φ webers linked with all the turns.
$e = N\dfrac{d\Phi}{dt} = L\dfrac{dI}{dt}$	Instantaneous e.m.f. generated in N turns linked by flux varying at a rate of $d\Phi/dt$ webers per sec.
$E = 4{\cdot}44\ \Phi_{max}Nf$	R.m.s. e.m.f. generated in N turns linked by Φ webers varying sinusoidally at f c/s.

Corresponding relationships in "mixed" units:

$F = \dfrac{4\pi IN}{10} = 1{\cdot}257\ IN$	F in gilberts
	I in amps
$H = \dfrac{1{\cdot}257\ IN}{l}$	H in oersteds
$B = \mu_s H$	l in cm
$\Phi = BA$	
$L = \dfrac{\Phi N}{I \times 10^8}$	B in gauss
	Φ in maxwells (" lines ")
$= \dfrac{1{\cdot}257\ \mu\ AN^2}{l \times 10^8}$	A in sq cm
$e = 10^{-8}N\dfrac{d\Phi}{dt} = L\dfrac{dI}{dt}$	L in henries
$E = 4{\cdot}44\ \Phi_{max}\ Nf \times 10^{-8}$	e and E in volts

14.9. TRANSFORMERS

Turns per volt, $\dfrac{N}{E} = \dfrac{1}{4{\cdot}44\ \Phi_{max} f}$ (Φ_{max} in webers)

or $\dfrac{10^8}{4{\cdot}44\ \Phi_{max} f}$ (Φ_{max} in maxwells)

Impedance Z_L in secondary circuit of ideal transformer is equivalent to n^2Z_L in the primary circuit, where n is the transformer ratio.

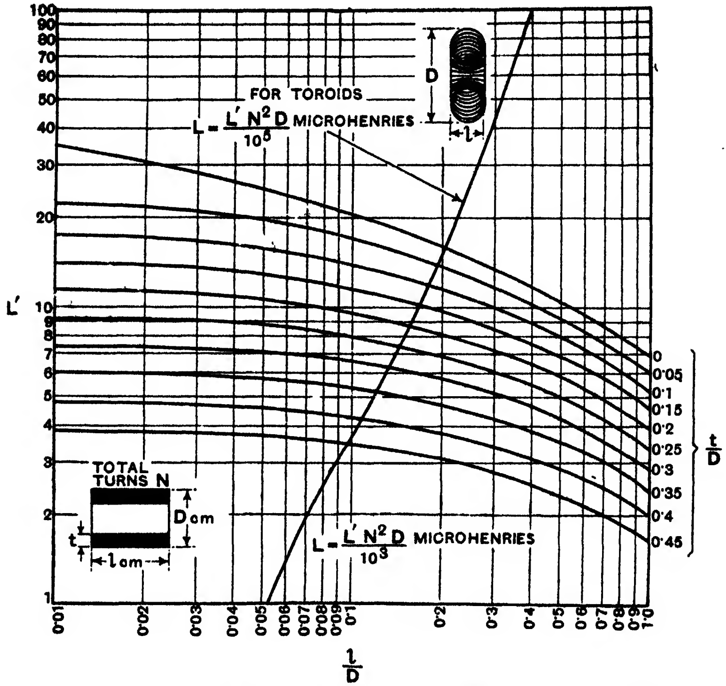


Fig. 14.5—Chart compiled from standard formulae by F. Charman, by which the self-inductance of coils of most shapes can be easily found

So where Z_L = load impedance

Z_P = desired impedance from primary side

the appropriate n is $\sqrt{Z_P/_LZ}$. Strictly, n is equal to $\sqrt{L_P/L_S}$, where L_P and L_S are the primary and secondary inductances, but in practice it is taken as the turns ratio.

Leakage inductance = $(1 - k^2)\, L_P$

where k = coefficient of coupling = $M/\sqrt{L_P L_S}$.

14.10. INDUCTANCE

The universal curve sheet, Fig. 14.5, covers the approximate calculation of single-layer solenoids and multi-layer coils of most shapes, including flat " pies ", and even toroids. From the curve and ordinate corresponding most closely to the dimensions of the particular coil, read off the factor L', which is then to be substituted in the formula given. A special curve is included for toroids, for which the dimensions have different meanings, as shown; and it should be noted that the formula differs by a factor of 0·01. The curve sheet enables the correct number of turns of wire on a former of given shape for a required inductance to be calculated. The tables in Sec. 14.40 then enable a suitable gauge of wire to be selected.

The following formula* for single-layer coils, which is accurate to within 1% when $l/D >$ about 0·4, fills in where the $t/D = 0$ curve leaves off, at $l/D = 1$:

$$L = \frac{N^2 D_{\text{cm}}}{101(l/D + 0{\cdot}45)} = \frac{N^2 D_{\text{in}}}{40(l/D + 0{\cdot}45)}$$

$$D_{\text{cm}} = \text{dia. in cm.}$$
$$D_{\text{in}} = \text{dia. in in.}$$

Coil Screening

The inductance of a coil screened by a non-magnetic cylindrical can

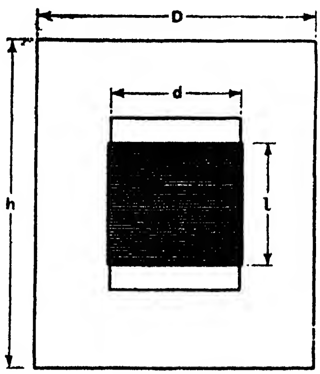


Fig. 14.6

is given approximately by multiplying the unscreened inductance by the following factor,† in which the dimensions are as shown in Fig. 14.6.

$$\frac{D^3 - d^3}{D^3}\left[1 - \left(\frac{l}{2h}\right)^2\right]$$

Mutual Inductance

The total inductance of two coils having self-inductance L_1 and L_2 and mutual inductance M is

$$L = L_1 + L_2 + 2M$$

Hence $$M = \tfrac{1}{2}(L - L_1 - L_2)$$

* Due to H. A. Wheeler, *Proc. I.R.E.*, Oct., 1928.
† Due to W. G. Hayman, *W.E.*, April, 1934.

This can be applied for calculating M when L, L_1, and L_2 are known or can be calculated; for example, two coaxial coils of equal diameter and turns per unit length:

$$M = \tfrac{1}{2}(L_{ad} + L_{bc} - L_{ac} - L_{bd})$$

L_{ad} is the inductance between a and d if the winding were con-

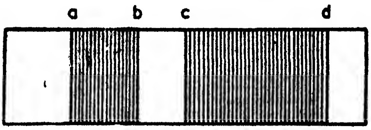


Fig. 14.7

tinuous between those points; and similarly for the others.

For more comprehensive data for calculation of L, M, and C, refer to F. Langford-Smith's *Radio Designer's Handbook* (Iliffe).

Inductive Transients

If L henries, carrying an initial current of I_0 amperes, is connected to

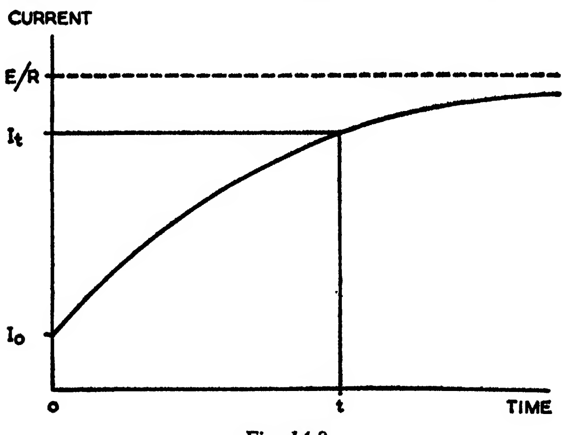


Fig. 14.8

a source of E volts in series with R ohms, the current after t seconds will be

$$I_t = \frac{E}{R} - \left(\frac{E}{R} - I_0\right) e^{-tR/L}$$

Alternatively
$$2{\cdot}30 \log_{10} \frac{E/R - I_t}{E/R - I_0} = -\frac{Rt}{L}$$

The quantity L/R is the *time constant*, equal to the time occupied by the first 63·2% (= 1 − 1/e) of the current change from I_0 to E/R.

Energy stored

The energy stored in L henries carrying I amps $= \frac{1}{2} I^2L$ joules.

14.11. ALTERNATING QUANTITIES

Values (for symbols, see p. 445).

When varying sinusoidally (i.e., with a sine waveform), the relative values, taking current as an example, are

$$i = I_{max} \sin \omega t \qquad (\omega = 2\pi f)$$

or, if the starting-point is the phase angle ϕ radians, $I_{max} \sin (\omega t + \phi)$

(Note: $\sin (\omega t + \pi/2) = \cos \omega t$)

$$I = \frac{I_{max}}{\sqrt{2}} \doteq 0{\cdot}707\ldots\ I_{max}$$

$$\text{So} \quad I_{max} = 1{\cdot}414\ldots\ I$$

$$\text{and} \quad I_{p-p} = 2{\cdot}828\ldots\ I$$

$$I_{av} = \frac{2}{\pi} I_{max} = 0{\cdot}637\ldots\ I_{max}$$

$$\text{So} \quad I = \frac{\pi}{2\sqrt{2}} I_{av} = 1{\cdot}11\ldots\ I_{av}$$

The ratio I/I_{av}, the *form factor*, varies with waveform. So metal-rectifier meters, which are usually calibrated in r.m.s. values though their readings are proportional to mean value, cannot be relied upon for accuracy unless the waveform is sinusoidal (Sec. 5.2). The same applies to most valve voltmeters, which depend on peak value or something between peak and mean (Sec. 5.14).

The mean value of any exclusively alternating quantity, taken over one or more whole cycles, is necessarily zero, but such statements as above are understood to refer to half cycles of rectified waveform.

Ref: " Values ", *W.W.*, Oct., 1946.

Addition

When more than one e.m.f. (alternating or direct) is acting in the same circuit, provided that the circuit is linear (i.e., current proportional to e.m.f.) the total current is equal to the sum of the currents due to each e.m.f. separately (Superposition theorem). This is true of instantaneous values, and also of r.m.s. values and peak values *if the waveforms are identical and directly in phase or opposition.* If not, the total power is equal to the sum of the separate powers, from which it follows that the total r.m.s. current (or e.m.f.) is equal to the square root of the sum of the squares, e.g.:

$$I = \sqrt{I_1^2 + I_2^2 + I_3^2} \text{ etc.}$$

Ref: " Total Power ", *W.W.*, March 1952.

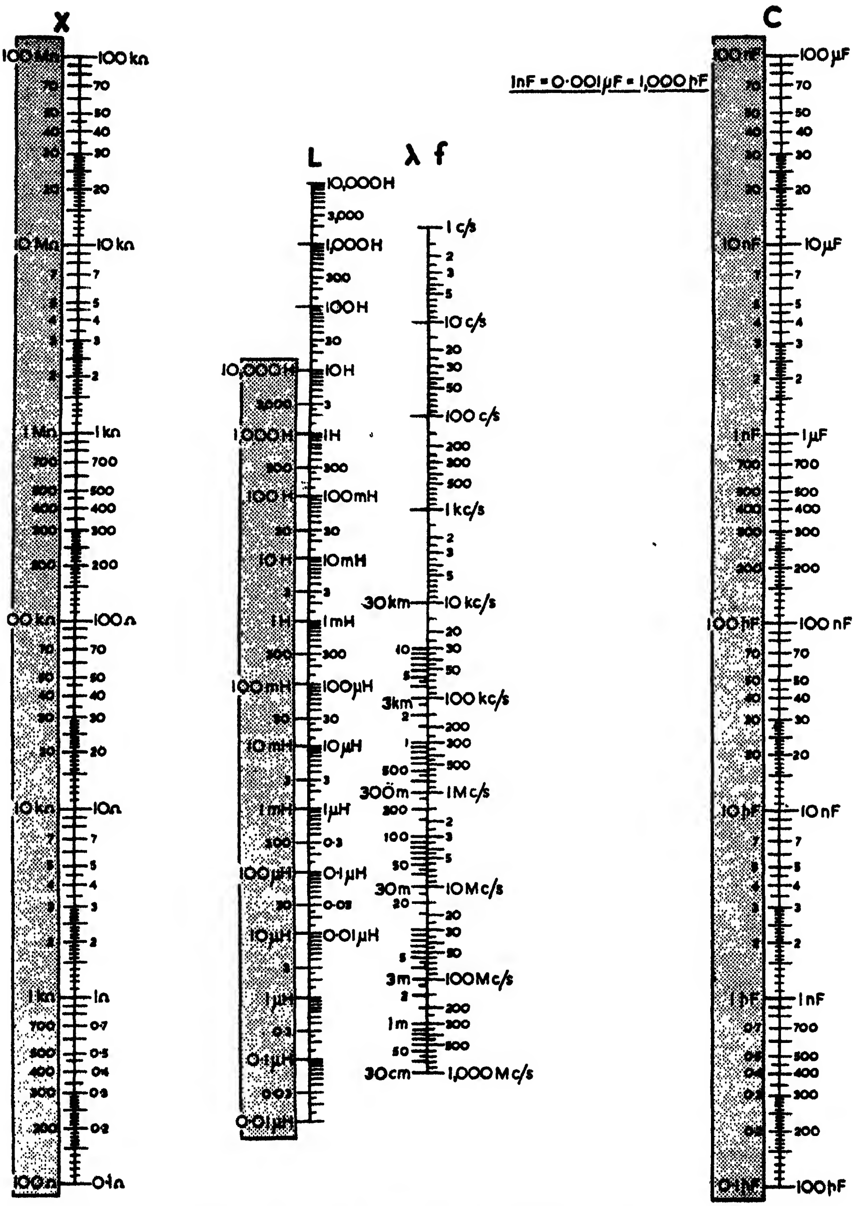


Fig. 14.9—This chart, used in the same way as Fig. 14.2, shows the reactance of a given inductance or capacitance at a given frequency, or the inductance and capacitance required to tune to a given frequency or wavelength, or vice versa. Use either all the shaded scales or none of them. E.g.: Find the L and C needed to give a dynamic resistance of 100 kΩ at 500 kc/s, assuming a Q of 200. The reactance must be $R/Q = 500\,\Omega$; connecting this with 500 kc/s shows L and C to be 160μH and 640 pF respectively

14.12. CALCULATION OF IMPEDANCES

Impedance, Z, is made up of resistance, R, and reactance, X, the latter being either positive (inductive), X_L, or negative (capacitive), X_C.

$$\text{So} \quad X = X_L + X_C$$
$$\text{where } X_L = \omega L \quad \text{and} \quad X_C = -1/\omega C \text{ (see Fig. 14.9)}$$
$$\text{and} \quad \omega = 2\pi f$$

Any system of resistances and reactances, however complicated, can be replaced by one resistance and one reactance, either in series or parallel as desired—but the values vary with frequency, and of course the series values differ from the parallel values. And the equivalence holds good only for steady-state currents—not transients. Most systems can be reduced to this simplest equivalent by successive use of the processes described below. It is easiest to work in reactances throughout, changing back to L and C if necessary by using the above formulae in reverse:

$$L = \frac{X}{\omega} \quad \text{and} \quad C = -\frac{1}{\omega X}$$

Incidentally, any capacitance can be expressed as a negative inductance, and vice versa.

Separate resistances *or* reactances in *series* can be combined by simple addition (remembering that X_C is negative). And so can inductances in series or capacitances in parallel.

X cannot be simply added to series R to give the impedance, for geometrically they are at right angles, as shown in Fig. 6.1, for example. One method is to use the Pythagorean relationship, thus:

$$Z^2 = R^2 + X^2, \text{ from which } Z = \sqrt{R^2 + X^2}$$

The phase angle, ϕ, can be calculated from

$$\tan \phi = X/R$$

The *reciprocals* of resistances, reactances or inductances in *parallel* (or of capacitances in series) can likewise be added. If they are turned round again into resistances, etc., it is convenient to remember that

$$\frac{1}{\frac{1}{a} + \frac{1}{b}} = \frac{ab}{a+b} \quad \text{and} \quad \frac{1}{\frac{1}{a} - \frac{1}{b}} = \frac{ab}{b-a}$$

But it is often simpler and clearer to reckon parallel circuits entirely in the reciprocal quantities G and B:*

$$G = 1/R \quad \text{and} \quad B = -1/X$$
$$B = B_C + B_L$$
$$\text{where } B_C = \omega C \quad \text{and} \quad B_L = -1/\omega L$$
$$Y = 1/Z = \sqrt{G^2 + B^2}$$

* "Admittance," *W.W.*, Jan. 1949.

To translate parallel circuit elements into their series equivalents, and vice versa, these relationships are very useful:

$$R_s = X_p \frac{R_p X_p}{R_p^2 + X_p^2} = \frac{R_p}{Q^2+1} \qquad X_s = R_p \frac{R_p X_p}{R_p^2 + X_p^2} = \frac{X_p}{1+1/Q^2}$$

$$R_p = \frac{R_s^2 + X_s^2}{R_s} = R_s(Q^2+1) \qquad X_p = \frac{R_s^2 + X_s^2}{X_s} = X_s(1+1/Q^2)$$

Calculations can sometimes be shortened by direct transfers from series R and X to parallel G and B, and vice versa:

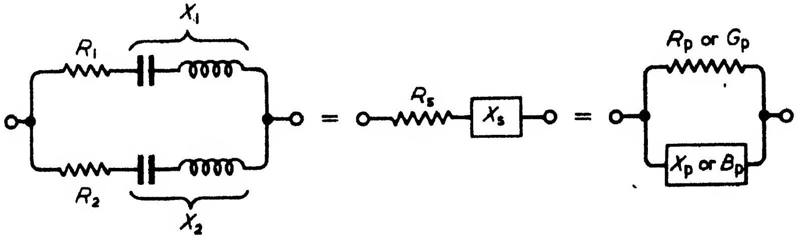


Fig. 14.10

$$G_p = \frac{R_s}{R_s^2 + X_s^2} = \frac{R_s}{Z_s^2} \qquad B_p = -\frac{X_s}{R_s^2 + X_s^2} = -\frac{X_s}{Z_s^2}$$

$$R_s = \frac{G_p}{G_p^2 + B_p^2} = \frac{G_p}{Y_p^2} \qquad X_s = -\frac{B_p}{G_p^2 + B_p^2} = -\frac{B_p}{Y_p^2}$$

A still more direct way is to use the following omnibus equivalent, which includes many oft-recurring combinations in which one or more of the values is zero, with consequent simplification:

$$R_s = \frac{R_1R_2(R_1+R_2)+R_1X_2^2+R_2X_1^2}{(R_1+R_2)^2+(X_1+X_2)^2} \qquad X_s = \frac{X_1X_2(X_1+X_2)+R_1^2X_2+R_2^2X_1}{(R_1+R_2)^2+(X_1+X_2)^2}$$

$$R_p = \frac{1}{G_p} = \frac{(R_1^2+X_1^2)(R_2^2+X_2^2)}{R_1(R_2^2+X_2^2)+R_2(R_1^2+X_1^2)}$$

$$X_p = -\frac{1}{B_p} = \frac{(R_1^2+X_1^2)(R_2^2+X_2^2)}{X_1(R_2^2+X_2^2)+X_2(R_1^2+X_1^2)}$$

Slide-rule Assistance

The most troublesome part of these calculations is the combination $R^2 + X^2$. There is, however, a simple method of doing it on a slide-rule. The four main scales of the rule will be referred to by the usual code, as A to D from top to bottom. Set the smaller of the two squared quantities on C against " 1 " on D. Set the cursor to the larger squared quantity on C. (If it is beyond " 10 " on C, set the smaller quantity to " 10 " on D instead of to " 1 ".) Note the cursor reading on A, and move the cursor to the right sufficiently to add 1 to that reading. The required value is then given by the cursor position on B.

Exactly the same procedure can be used to calculate Z (or Y) from R and X (or G and B), except that the answer is read off C, that being of course the square root of the figure on B. With a little practice either kind of calculation can be done very quickly indeed, but of course one must be careful to get the decimal point right, by a rough mental check.

Refs: "Impedance and Admittance Calculations," by F. Oakes and E. W. Lawson. *W.W.*, July, 1955.

The Modern Slide Rule, 3rd edn., by R. Stender and K. K. McKelvey. (Cleaver-Hume, 1960.)

The j Method

Where it is necessary to calculate the impedance of a circuit in general, using letter symbols, doing so by means of the foregoing formulae would soon become very involved. For this and much other work it is simpler to use the operator j.

The subject is covered in many books,* but briefly the symbol j means that a vector representing the quantity to which it is prefixed

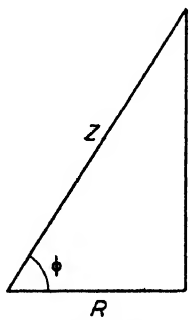


Fig. 14.11

is to be turned anticlockwise through a right angle. So the complex impedance operator

$$Z = R + jX$$

can be interpreted as an instruction to move R steps to the right and X steps upward, the distance from the starting point then being Z steps, as in Fig. 14.11. The phase angle is given by $\tan \phi = X/R$.† The usefulness of j appears when a number of impedances have to be combined. If this is done in the $\sqrt{(R^2 + X^2)}$ form the algebra soon

* E.g., *Basic Mathematics for Radio and Electronics Students*, by F. M. Colebrook and J. W. Head (Iliffe); *A Mathematical Tool Kit for Engineers*, 2nd edn., by H. A. Webb and D. E. Ashwell (Longmans, 1959); and *A.C. Network Analysis by Symbolic Algebra*, by W. H. Miller (Classifax Publications). See also "j", *W.W.*, Feb., 1948.

† It is necessary, of course, to have regard for the sign of X, and possibly of R.

becomes very cumbersome; the j form, with $j = \sqrt{-1}$, is much simpler. (The word "complex", which is the name for this kind of algebra, should not be taken in its usual sense; it means quantities which include both ordinary ("real") and $\sqrt{-1}$ ("imaginary") numbers. An important peculiarity is that all terms with j can be added together, as can those without j, but the two kinds must be kept separate. Thus

$$(p + jq) + (r + js) = (p + r) + j(q + s)$$

Here p and r represent resistances (or conductances), and q and s represent reactances (or susceptances). In multiplication, $j^2 = -1$, so

$$(p + jq)(r + js) = (pr - qs) + j(qr + ps)$$

The first term $(pr - qs)$ represents the resistance, and the second $(qr + ps)$ the reactance. Also

$$\frac{1}{p + jq} = \frac{p - jq}{p^2 + q^2}$$

$$\frac{p + jq}{r + js} = \frac{(pr + qs) + j(qr - ps)}{r^2 + s^2}$$

$$\sqrt{(p + jq)} = \sqrt{\tfrac{1}{2}(\sqrt{p^2 + q^2} + p)} + j\sqrt{\tfrac{1}{2}(\sqrt{p^2 + q^2} - p)}$$

Ref: "The Complex Number". *W.W.*, Feb. 1953.

14.13. DUALS

It may have been noticed that of the two pairs of equations half way down p. 466 each pair is exactly the same as the other, except that the following have changed places:

R and *G*

X and *B*

series and parallel

G, *B*, and parallel are said to be *duals* of *R*, *X*, and series respectively; and vice versa. A relationship between one set having been established, the corresponding relationship for the other set can be written down by substituting the duals. Other duals are:

I and *V*

Z and *Y*

C and *L*

Therefore, given that $X = \omega L$, one knows at once that $B = \omega C$; and so on. This is a useful means of extending a list of standard formulae.

Ref: "Duals". *W.W.*, April 1952.

14.14. Y–Δ OR T–Π TRANSFORMATION

A very common circuit formation is the star, Y, or T; and another is the delta (Δ) or Π. Filters and attenuators (Secs. 14.24 to

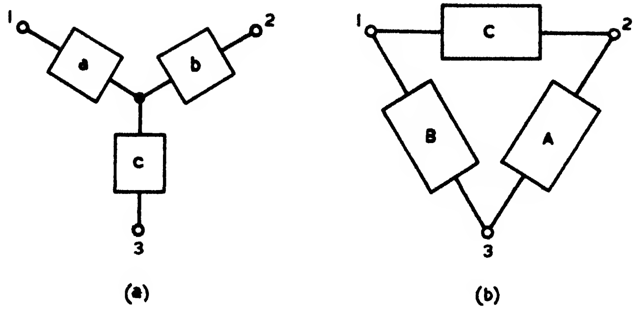


Fig. 14.12—(a) *Star* (⅄) *or T.* (b) *Delta* (Δ) *or* Π

14.26) are examples. The impedance values for a ⅄ equivalent to a given Δ (or vice versa) can be calculated from the following equations, in which the impedances are complex (i.e., $R + jX$).

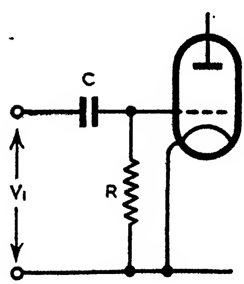
Δ *to* ⅄	⅄ *to* Δ
$a = \dfrac{BC}{A+B+C}$	$A = \dfrac{ab+bc+ca}{a}$
$b = \dfrac{CA}{A+B+C}$	$B = \dfrac{ab+bc+ca}{b}$
$c = \dfrac{AB}{A+B+C}$	$C = \dfrac{ab+bc+ca}{c}$

Ref.: "Pie Tea," *W.W.*, Oct. 1956.

14.15. FREQUENCY CUT-OFF CURVES

One of the commonest circuit calculations is finding (for example) the proportion of the input voltage reaching the valve in Fig. 14.13.

Fig. 14.13—Example of resistance-reactance potential divider



Here C and R form a potential divider across the supply, and what is required is the ratio of R to the impedance comprising C and R. The required ratio is $R/\sqrt{(R^2 + X^2)}$, and calculating points for plotting numerous frequency curves is tedious. But putting this expression in

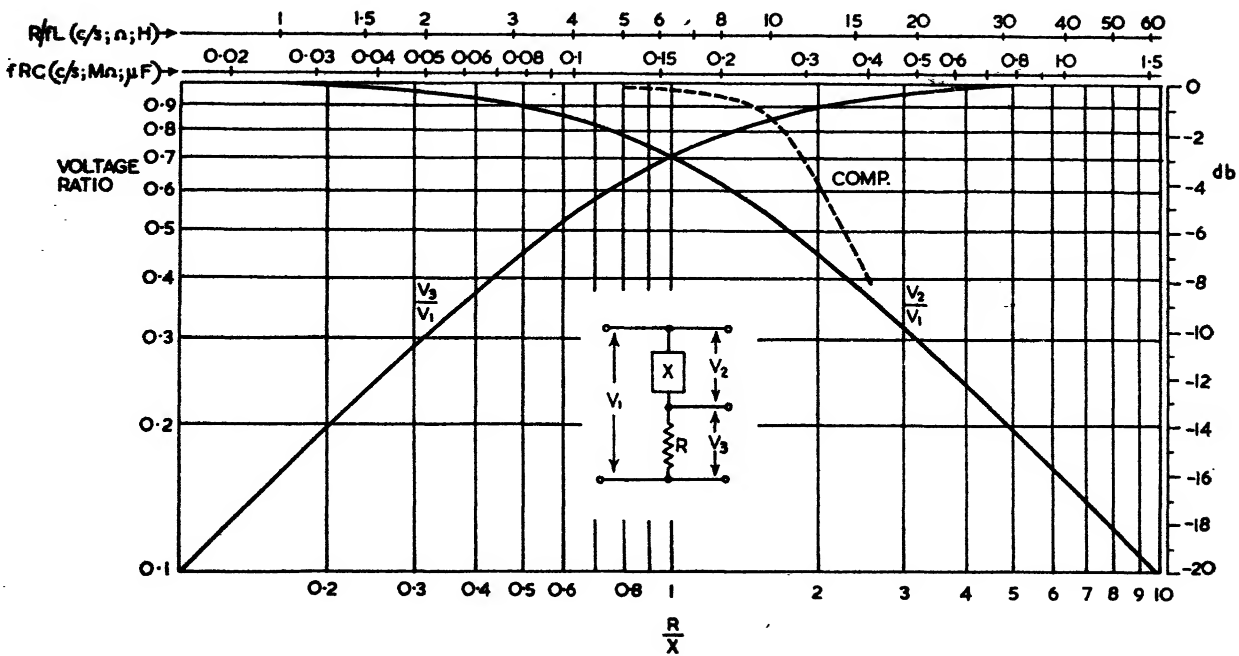


Fig. 14.14—Universal curves whereby the division of voltage between a resistance and a reactance can be read off. This is particularly useful in estimating the performance, at different frequencies, of valve amplifier circuits. The dotted curve refers to the inductance-compensated circuit (Fig. 14.15) in which $L = 0.414\ R^2C_s$ and $C = 0.35\ C_s$ (Cocking's Television Receiving Equipment, *4th edn., p. 213)*

the form $1/\sqrt{[(X/R)^2 + 1]}$ makes clear that the voltage ratio is a function of X/R only, so if plotted with suitable scales, as in Fig. 14.14, a single curve suffices.

E.g: $R = 0{\cdot}25\ \text{M}\Omega$, $C = 5\ \text{nF}$ ($= 0{\cdot}005\ \mu\text{F}$); find loss at 100 c/s.

From Fig. 14.9, X at 100 c/s $= 0{\cdot}318\ \text{M}\Omega$. So $R/X = 0{\cdot}785$. From the V_3/V_1 curve in Fig. 14.14 the voltage across R relative to the input voltage is 0·62, or about 4 dB down. The relative voltage across C, from the V_2/V_1 curve, is 0·78, or 2 dB down. Alternatively, using the fRC scale, $RC = 0{\cdot}00125$ or 1/800, so if the fRC scale is multiplied by 800 it becomes a frequency scale for both curves.

The topmost scale can be used similarly for resistance-inductance combinations.

It is useful to remember that:

(1) When $X = R$ the loss across either R or X is 3 dB ($=1/\sqrt{2}$ voltage ratio).
(2) The loss, taking output from across R, when f = reciprocal of time constant (i.e., $1/RC$ or R/L), is 16 dB.
(3) The slope of the curve from about 10 dB downwards is 6 dB per octave (i.e., doubling or halving of frequency).

From these three data, a rough frequency curve can be plotted.

The standard shape depends on the input voltage being constant for all frequencies; but this condition can usually be satisfied by including the generator resistance in the circuit.

The dual principle enables the same curve and calculations to be used for G and B in parallel, with constant current input. This condition very nearly holds in pentode output circuits and can be made to hold exactly by adding $1/r_a$ to G.

The whole subject of curves for this and for more complicated circuits is treated at length by N. H. Crowhurst.*

14.16. RESONANCE

Resonance in a circuit containing L, C and R is defined in various ways, as for example by zero phase difference between e.m.f. and current, or by maximum current or voltage in various parts of the circuit; and the frequencies at which these conditions are obtained may differ according to what is varied in order to obtain them. A summary of these conditions, though concisely set out, occupies four whole pages in the *Army Handbook of Line Communication*, Vol. 1 (H.M.S.O.), pp. 227–230. When Q (Sec. 14.17) is reasonably large, say >10, the differences between these conditions are for most purposes negligible, but for accurate measurements—especially with

* "The Prediction of Audio-frequency Response", *E.E.*, Nov. and Dec. 1951, Jan. and Feb. 1952.

low Q—it is necessary to take care to use the correct equations. There is room here for only a few of the most important.

Series Resonance (Fig. 14.16*a*) takes place when $X_L = X_C$, which reduces the total impedance to r. If f_r denotes the frequency of resonance, and $\omega_r = 2\pi f_r$:

$$\omega_r L = \frac{1}{\omega_r C} \quad \text{and} \quad \omega_r^2 = \frac{1}{LC} \quad \text{and} \quad f_r = \frac{1}{2\pi\sqrt{LC}}$$

This can be put into still other forms to suit the most convenient working units of L and C; e.g:

$$f_r = 159{\cdot}2/\sqrt{LC} \qquad \text{[kc/s; μH; μF] or [Mc/s; μH; pF]}$$

A slide rule having a reciprocal scale can be used to evaluate this

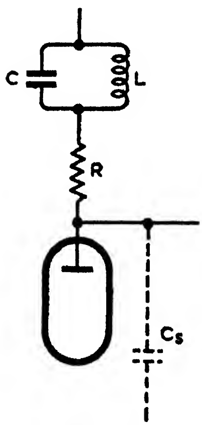


Fig. 14.15

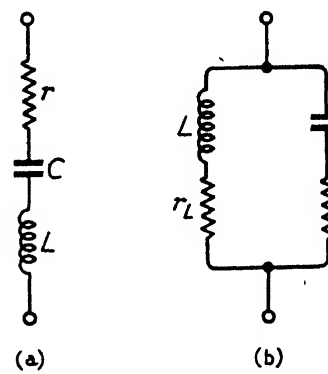


Fig. 14.16

equation in one setting (F. Oakes, *W.W.*, Sept. 1953). Set L on scale B against 253 on the left-hand half of A, and read f on the reciprocal scale of the slide against C on A. If the first significant figure of L and/or C multiplies an even power of 10 (e.g., 256 pF or 18 mH), use the left-hand half of A or B; if an odd power (e.g., 2·5 nF or 65 μH), use the right-hand half.

If wavelengths are preferred to frequencies:

$$\lambda_r = 1{\cdot}885\sqrt{LC} \qquad \text{[μH; pF]}$$

The above condition gives zero phase difference, and minimum impedance on varying L, C or f (r assumed constant). But it does not give maximum voltage across C on varying either C or f. Assuming constant applied voltage, the frequency for this voltage resonance depends on whether r or Q is assumed constant; in general, the frequency differs from f_r by a fraction of the order of $\pm 1/Q^2$.

Parallel Resonance (Fig. 14.16*b*) is more complicated, even when r_C is neglected, because the circuit as shown (which approximately represents the behaviour of an actual tuned circuit) is partly series and partly parallel. If the values given are converted into their parallel equivalents (Sec. 14.12), these values can be used in the simple formulae given above for series resonance; but with fixed values this structure does not very well represent the behaviour of a tuned circuit when frequency varies. With high Q, the series and parallel values of L and C are almost identical, but the parallel resistance (" dynamic resistance "), denoted by R, is as many times greater than X_L and X_C as these are greater than r:

$$R \simeq \frac{L}{rC} \simeq \omega_r LQ \qquad [Q \not< \text{about } 10]$$

For this purpose, $r = r_L + r_C$.

The exact equation for zero phase difference, r_C being neglected, and either r_L or Q assumed constant, is

$$\omega^2 = \frac{1}{LC} - \frac{r_L^2}{L^2} = \omega_r^2\left(1 - \frac{1}{Q^2}\right) \qquad (\omega_r \text{ as for series resonance})$$

This also gives maximum impedance on varying C, and therefore maximum voltage across the circuit, assuming constant current through it.

If either arm of a parallel tuned circuit is tapped (Fig. 14.17), the

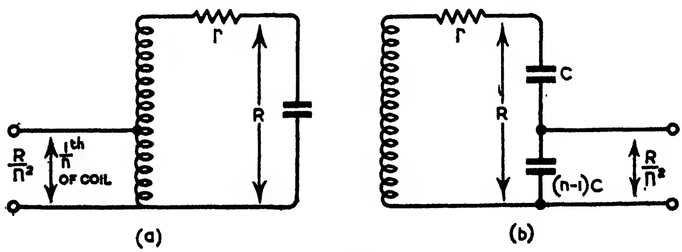


Fig. 14.17

impedance across the tapped-off portion at resonance is still resistive and is proportional to the square of the tapping ratio.

Ref: " Resonance Curves ", *W.W.*, Jan. 1953.

14.17. Q

The basic definition of the Q of a circuit is:

$$\frac{2\pi \times \text{energy stored}}{\text{energy dissipated}}$$

in the circuit per half-cycle.

This is the only definition applicable to circuits in which L, C and

r are distributed; e.g., transmission lines and resonant cavities. Where they are lumped, this definition amounts to

$$Q = X/r$$

where X is the reactance of one kind and r is the series resistance covering all losses. This X/r ratio is also equal to the ratio of V, the voltage across the whole reactance of either kind in a circuit at series resonance, to E, the e.m.f. injected in series. This ratio is also

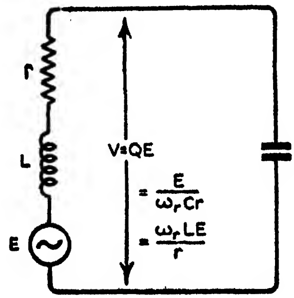


Fig. 14.18

known as the circuit magnification, m. But when, as is usual, resonance is judged by the maximum parallel voltage, there is a discrepancy between m and Q, which is negligible unless $Q <$ about 5. If L and r (Fig. 14.18) are shunted by self-capacitance C_0, as in the type of Q meter shown in Fig. 7.19, the measured V/E is the apparent Q, or Q':

$$Q' = Q\frac{C}{C + C_0}$$

Except for the usually negligible discrepancy just mentioned, m is the same as Q', and therefore in practice differs from the true Q. Certain other methods of measurement (e.g., Sec. 9.28) give true Q, and these methods also correspond with the conditions under which tuned circuits are most commonly used.

An average value of Q for tuned circuits is about 100. If a resonant circuit includes various items in parallel—coil, tuning, capacitor, valve holder, wiring, etc.—each with its own Q and C, then the Q of the whole circuit is given by

$$\frac{1}{Q} = \frac{1}{Q_L} + \frac{C_1}{Q_1 C} + \frac{C_2}{Q_2 C} + \text{etc.},$$

where C is the total capacitance of the circuit, Q_L is the Q of the coil as such, and the numbered quantities refer to each separate item having capacitance (including the self-capacitance of the coil regarded as a separate entity).

Refs: " The Development of Q-meter Methods of Impedance Measurement ", by A. J. Biggs and J. E. Houldin. *Proc. I.E.E.*, Part III, July 1949.
" Q-meter Controversy " (Appendix), by P.H. *W.W.*, June 1949, p. 217.
" Q ". *W.W.*, July 1949.

14.18. TUNING CURVES

The response curve around the resonant frequency is of special interest. Its shape, for a single series circuit with constant applied e.m.f., or parallel circuit with constant current, is practically the same as those in Fig. 14.14. It is only when the frequency off tune, f', is not very much less than the frequency of resonance, f_r, that the shapes begin to differ appreciably. A tuned circuit driven by a pentode (Fig. 14.19) is a close approximation to constant signal-current conditions. The V_2/V_1 curve in Fig. 14.14 is converted into a resonance or selectivity curve by exchanging the R/fL scale for $f'L/r$ or Qf'/f_r as in Fig. 14.20. The dB scales are the same, but it is perhaps more convenient to have as the alternative voltage-ratio scale the reciprocal of that in Fig. 14.14, denoted by S. For example, suppose Q is 80 and f_r is 500 kc/s. Then by multiplying the horizontal scale by 500/80 it becomes a scale of " kc/s off tune " for that particular

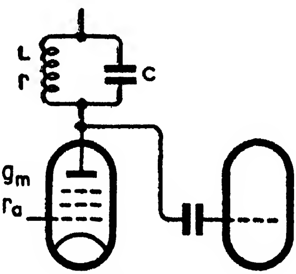


Fig. 14.19

circuit. To find the adjacent-channel (9 kc/s off tune) selectivity, the scale reading is $80 \times 9/500 = 1{\cdot}44$, and the single-circuit curve shows the selectivity S to be 3, meaning that when the circuit is detuned by 9 kc/s the signal voltage needed to maintain the response as before is 3 times as great; or that the adjacent-channel response is one-third of that at resonance, or nearly 10 dB down. Two circuits entirely uncoupled except by a perfectly screened valve give a ratio of $3^2 = 9$; three, of 27; and so on.

The equation for calculating these curves is

$$S \simeq (4\alpha^2 + 1)^{N/2} \qquad [f' << f_r]$$

where $\alpha = Qf'/f_r = 2\pi f'L/r$ $\quad N$ = number of tuned circuits
f_r = frequency of resonance, $\quad f'$ = difference between f_r and actual frequency

When $N = 1$, the formula is $\sqrt{4\alpha^2 + 1}$.
Maximum response (at resonance), $m_{max} = Q$.
If in a stage of amplification $r_a >> R(= \omega LQ)$, the voltage gain of the stage, $A_1 \simeq Qg_m\omega L = Qg_m/\omega C = Q^2 g_m r = g_m R$.

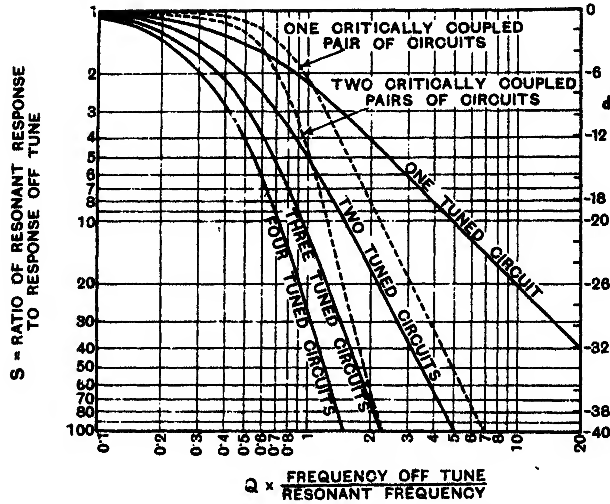


Fig. 14.20—Universal response curves, showing the extent to which the response falls off as there is departure from the resonant frequency. Comparing one critically-coupled pair of circuits with two single circuits, one can see that the steepness of the slope well away from resonance is the same, but the coupled pair gives a more level response near resonance

Coupled Circuits (Fig. 14.21)

Coefficient of coupling, k, $= X_m/\sqrt{X_1X_2}$, where X_1 and X_2 are the reactances of the two circuits and X_m is the mutual reactance. For inductive coupling

$$k = M/\sqrt{L_1L_2}$$

The selectivity equation for one pair of coupled circuits is

$$S \simeq \frac{2}{1+\beta^2}\sqrt{4\alpha^4 + 2(1-\beta^2)\alpha^2 + (1+\beta^2)^2/4}$$

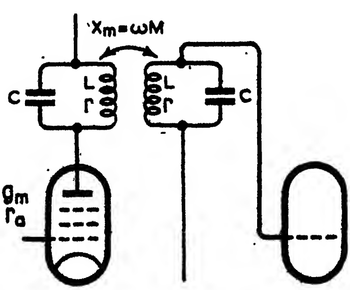


Fig. 14.21

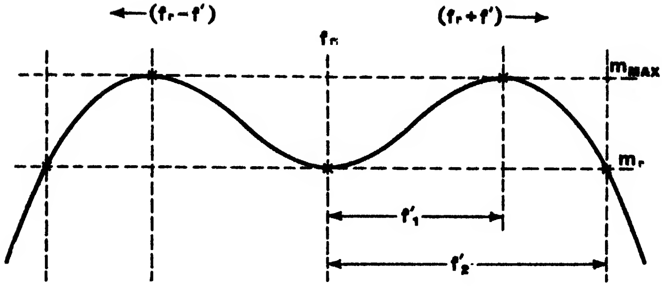


Fig. 14.22

where $\beta = Qk = X_m/r = QM/L$ when coupling is wholly inductive. Other conditions as for single tuned circuits.

Condition 1 $\quad \beta = 1$ (critical or optimum coupling)
$X_m = r$

Then m_{max} at resonance $= Q/2$

$$\text{and } A \simeq Qg_m\omega L/2 \qquad [r_a >> R]$$

$$S \simeq \sqrt{4\alpha^4 + 1},$$

or, for N pairs of circuits, $(4\alpha^4 + 1)^{N/2}$, from which response curves can be worked out (dotted in Fig. 14.20).

Condition 2 $\quad \beta < 1$ (under-coupling)
$m_{max} < Q/2$

Condition 3 $\quad \beta > 1$ (over-coupling) (Fig. 14.22)
$m_{max} = Q/2$, but not at resonance

$$\text{Response at resonance, } m_r, = \frac{Q\beta}{1 + \beta^2} \quad \text{and} \quad \frac{m_{max}}{m_r} = \frac{1 + \beta^2}{2\beta}$$

$$f_1' = \pm \frac{f_r\sqrt{\beta^2 - 1}}{2Q}$$

$$f_2' = \sqrt{2}f_1'$$

from which five important points can easily be calculated:

$$A \text{ (at resonance)} \simeq \frac{Qg_m\omega L\beta}{1 + \beta^2}$$

$$A \text{ (at peaks)} \simeq \frac{Qg_m\omega L}{2}$$

$$S \text{ at resonance} = (1 + \beta^2)/2\beta$$

14.19. MILLER EFFECT

The effect of the interelectrode capacitances (Fig. 14.23) on the input is equivalent to a capacitance C_{in} in parallel with a resistance R_{in}.

$$C_{in} = C_{gk} + C_{ga}(1 + A \cos \theta)$$
$$R_{in} = -1/(\omega C_{ga} A \sin \theta)$$

[R_{in} is positive with capacitive loads ; negative with inductive loads]

where A = stage gain, V_a/V_g

θ = angle by which V_a leads $-\mu V_g$

[θ is positive with inductive load]

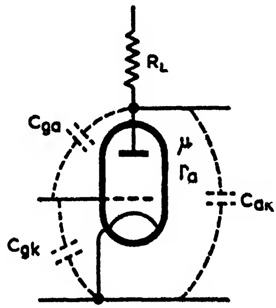


Fig. 14.23

With resistive load (neglecting, or neutralizing by inductance, the capacitance C_{ak}):

$$C_{in} = C_{gk} + C_{ga}\left(1 + \frac{\mu R_L}{r_a + R_L}\right)$$

Each capacitance denoted includes capacitance inside and outside the valve.

14.20. NEGATIVE FEEDBACK

Effect on Amplification

If a proportion, B, of the output voltage of an amplifier giving an amplification A, is fed back to the input, the overall amplification is altered to

$$A' = \frac{A}{1 - AB}$$

In general, distortion and noise arising in the amplifier are reduced in the same ratio (but see below regarding hum).

Suppose the voltage amplification is 500, and 6% of the output is fed back in opposition ($B = -0{\cdot}06$). Then $A' = \frac{500}{1 + 30} = 16$.

If $-AB >> 1$, $A' \simeq \frac{1}{-B}$; that is to say, the amplification is virtually unaffected by minor changes in the amplifier itself and depends only on the feedback circuit.

In general, A and B are complex; i.e., their phase angles are not exactly 0° and 180° respectively. If at any frequency the overall phase angle $= 0$ (positive feedback) and $AB \nless 1$, the amplifier is unstable and may oscillate at that frequency.

Effect on Output Resistance

The apparent r_a of the valve may be either increased or decreased, depending on whether current or voltage feedback is used. For example, negative feedback due to an un-bypassed cathode resistor is proportional to the signal *current* through it. Feedback taken from a transformer or potential divider across the load, and therefore proportional to the output *voltage*, reduces the apparent r_a to

$$r_a' = \frac{r_a}{1 - \mu B}$$

So the r_a of a high-μ valve such as a tetrode or pentode is reduced much more by feedback than the amplification.

Cathode Follower

In a cathode follower, $B = -1$, $r_a' = r_a/(\mu + 1)$, and if 1 is neglected in comparison with μ this reduces to

$$r_a' \simeq 1/g_m$$

If g_m is, say, 10 mA/V, r_a' is rather less than 100 Ω, so a cathode follower can be used to feed quite a low load impedance direct. It acts as a current or power amplifier; the voltage gain < 1:

$$A' = A/(A + 1) \simeq \frac{R_L}{R_L + 1/g_m}$$

where R_L = load resistance.

Hum (see also Sec. 14.25)

Although negative feedback reduces noise arising within the feedback loop, it may either reduce or increase hum resulting from ripple in the supply voltage, depending on the type of circuit.* Feeding back direct from the anode of a transformer-coupled output valve is liable to increase hum, perhaps very greatly. Feedback from the secondary of the transformer, or from a parallel-feed, reduces hum.

14.21. VALVE EQUIVALENT GENERATOR

Fig. 14.24*a* shows the essentials of a valve amplifier circuit, and *b* the equivalent voltage generator, assuming linear valve characteristics. In order to conform to the usual convention of reckoning anode and grid voltages relative to cathode, it is necessary for the generator

* "Negative Feedback and Hum". *W.W.*, July, 1961.

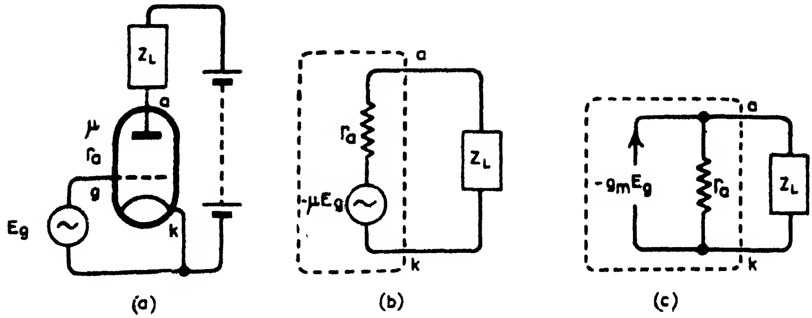


Fig. 14.24

voltage to have a negative sign. *c* is the equivalent current generator, which is more suitable than *b* for simulating a pentode, but both give the same results. The current-generator branch is assumed to have infinite internal resistance, so as not to short-circuit r_a and R_L.

Refs.: *W.W.*, April 1947, pp. 129–130, and April 1951, pp. 152–4.

14.22. TRANSISTOR EQUIVALENT GENERATOR

The subject of transistor equivalent circuits is much more complicated than that of valves, for two main reasons: (1) common impedance causes feedback even at z.f., and (2) high-frequency effects are, in general, appreciable at much lower frequencies—even at 1 kc/s in some types. Many varieties of equivalent circuit are used, some with one and some with two generators. A single-generator circuit, with capacitors approximately simulating high-frequency behaviour, is shown as Fig. 14.1. Fig. 14.25 shows the simple low-frequency T networks for (*a*) common-emitter and (*b*) common-base configurations. In each case both voltage and current generators are shown, but these are alternatives, to be included only one at a time. Note that the outputs of (*a*) and (*b*) are opposite in phase. α_b is what is often denoted by α, and α_e is often denoted by α' or β.

$$\alpha_e = \frac{\alpha_b}{1 - \alpha_b} \qquad r_{ce} = r_{cb}(1 - \alpha_b)$$

$$\alpha_b = \frac{\alpha_e}{1 + \alpha_e} \qquad r_m = \alpha_b r_{cb}$$

For many purposes the much simpler common-emitter equivalent circuit, Fig. 14.25 (*c*) is sufficiently approximate. This is so when the internal feedback is either relatively insignificant or its effect is neutralized. There is, however, the practical difficulty that the mutual conductance, g_m, is much less linear with respect to signal amplitude than α.

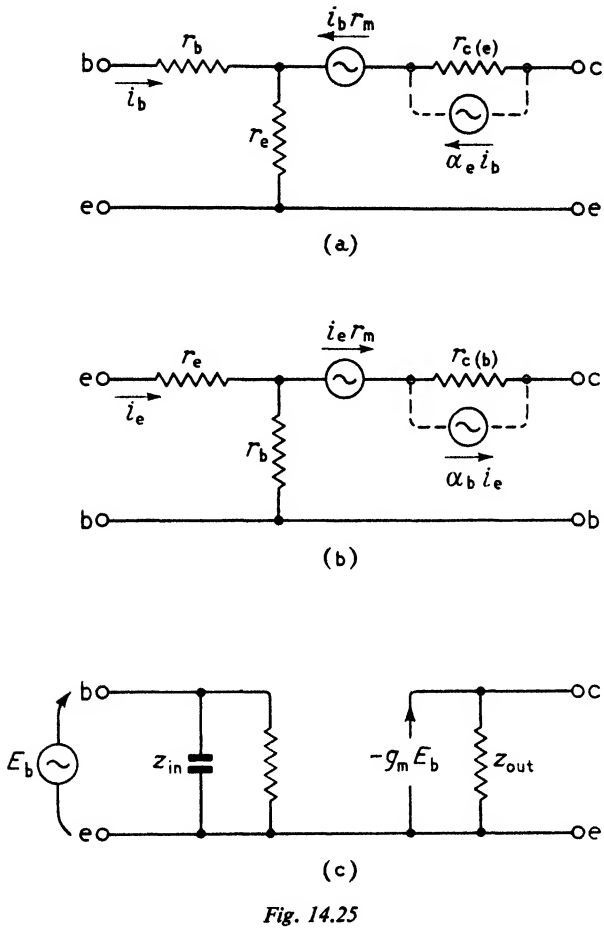


Fig. 14.25

14.23. THÉVENIN'S THEOREM

This is the general principle, of which the valve equivalent generator is a simple example. It says that if you connect any impedance Z to any two points of a linear circuit, the current I you will get through Z will be the same as if the circuit were a generator of an e.m.f. equal to the voltage V_0 between the two points when Z is not connected, in series with an impedance Z_0 equal to that measured between the two points, all circuit e.m.fs having been reduced to zero; i.e:

$$I = \frac{V_0}{Z + Z_0}$$

Fig. 14.26 is an example of a circuit network that cannot be reduced to a simple equivalent by successive combining of series and parallel elements; and solving the seven simultaneous Kirchhoff equations is tedious. It is much easier to find I by applying Thévenin. The open-circuit voltage between a and b is quite easy to calculate, and so is the " generator " resistance (R_1 and R_2 in parallel, in series with R_3 and R_4 in parallel).

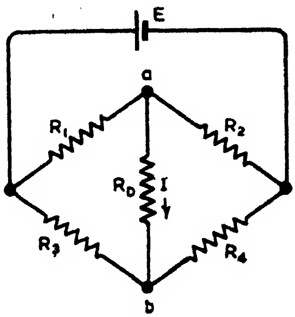


Fig. 14.26

A chart, based on Thévenin's theorem, for determining the pair of preferred-value resistors needed to make a specified potential divider, has been provided by J. Willis (*W.W.*, Sept. 1958).

The dual form (Sec. 14.13) of the theorem can be used to find the voltage V that would occur across a load admittance Y, in terms of the generator admittance Y_0 and the *short*-circuit current I_0 between the load terminals:

$$V = \frac{I_0}{Y + Y_0}$$

Although Thévenin's theorem was originally restricted to steady-state conditions, it can also be applied to calculating transients.*

Ref: " Thévenin's Theorem ". *W.W.*, March 1949.

14.24. ATTENUATORS

Table 14.10 covers the most important types of attenuator used between equal resistances, R (Fig. 14.27). The required attenuator resistances are equal to R multiplied by the r values given in the Table. Consider, for instance, a 600-Ω 3-dB T: the appropriate columns are r_1 and r_2, which at 3 dB are 0·171 and 2·84 respectively. Multiplied by 600 these give R_1 and R_2 as 102·6 Ω and 1,704 Ω. If one form of attenuator would require excessively low or high resistance values, try another.

* "The Extended Employment of Thévenin's Theorem", by A. Lee and D. K. C. MacDonald. *W.E.*, Nov. 1945.

Table 14.10

dB loss	$a = V_{in}/V_{out}$	r_1	r_2	r_3	r_4	r_5	r_6
0	1	0	∞	0	∞	0	∞
0·1	1·012	0·00576	86·9	0·0115	174	0·0116	86·4
0·2	1·023	0·0115	43·4	0·0230	86·9	0·0233	42·9
0·3	1·035	0·0173	28·9	0·0345	57·9	0·0351	28·5
0·4	1·047	0·0230	21·7	0·0461	43·4	0·0471	21·2
0·5	1·059	0·0288	17·4	0·0576	34·8	0·0593	16·9
0·6	1·072	0·0345	14·5	0·0691	29·0	0·0715	14·0
0·8	1·096	0·0460	10·8	0·0922	21·7	0·0965	10·36
1·0	1·122	0·0575	8·67	0·115	17·4	0·122	8·20
1·5	1·188	0·0861	5·76	0·174	11·6	0·188	5·30
2	1·259	0·115	4·30	0·232	8·72	0·259	3·86
3	1·413	0·171	2·84	0·352	5·85	0·413	2·42
4	1·585	0·226	2·10	0·477	4·42	0·585	1·71
5	1·778	0·280	1·64	0·608	3·57	0·778	1·28
6	1·995	0·332	1·34	0·747	3·01	0·995	1·005
7	2·239	0·382	1·12	0·896	2·61	1·24	0·807
8	2·512	0·431	0·946	1·057	2·32	1·51	0·661
9	2·818	0·476	0·812	1·23	2·10	1·82	0·550
10	3·162	0·520	0·703	1·43	1·92	2·16	0·462
12	3·98	0·598	0·536	1·86	1·67	2·98	0·335
14	5·01	0·667	0·416	2·41	1·50	4·01	0·249
15	5·62	0·698	0·367	2·72	1·43	4·62	0·216
16	6·31	0·726	0·325	3·08	1·38	5·31	0·188
18	7·94	0·776	0·256	3·91	1·29	6·94	0·144
20	10·00	0·818	0·202	4·95	1·22	9·00	0·111
25	17·78	0·894	0·113	8·86	1·12	16·8	0·0596
30	31·62	0·939	0·0633	15·8	1·07	30·6	0·0327
35	56·2	0·965	0·0356	28·1	1·04	55·2	0·0181
40	100·0	0·980	0·0200	50·0	1·02	105	0·0101
45	177·8	0·989	0·0112	88·9	1·011	177	0·00566
50	316·2	0·994	0·00632	158	1·006	315	0·00317
55	562	0·996	0·00356	281	1·0036	561	0·00178
60	1000	0·998	0·00200	500	1·0020	999	0·00100

Ref: "Designing Resistive Attenuating Networks", by P. K. McElroy. *Proc. I.R.E.*, March 1935.

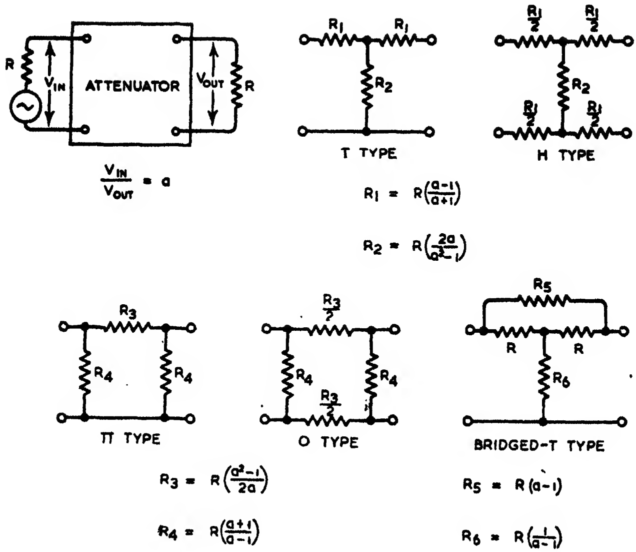


Fig. 14.27

14.25. SMOOTHING AND DECOUPLING FILTERS

If the load impedance $>> 1/\omega C$, the simple RC smoother or decoupler (Fig. 14.28) is an example of Sec. 14.15. In Fig. 14.29 the same information (full-line curves) is presented in convenient form. In Fig. 14.28, V_{in} and V_{out} refer to the ripple or hum voltage. Then

$$\frac{V_{\text{in}}}{V_{\text{out}}} = \sqrt{\left(\frac{R}{X_C}\right)^2 + 1}.$$

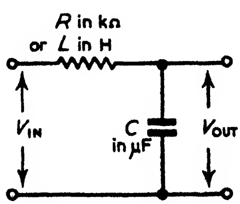


Fig. 14.28

If, as would normally be so, R/X_C is at least 4,

$$\frac{V_{\text{in}}}{V_{\text{out}}} \simeq \frac{R}{X_C} = 2\pi fRC \qquad [R \nless 4X_C, \text{ or } fRC \text{ in c/s–k}\Omega\text{–}\mu\text{F} \nless \text{ say } 600]$$

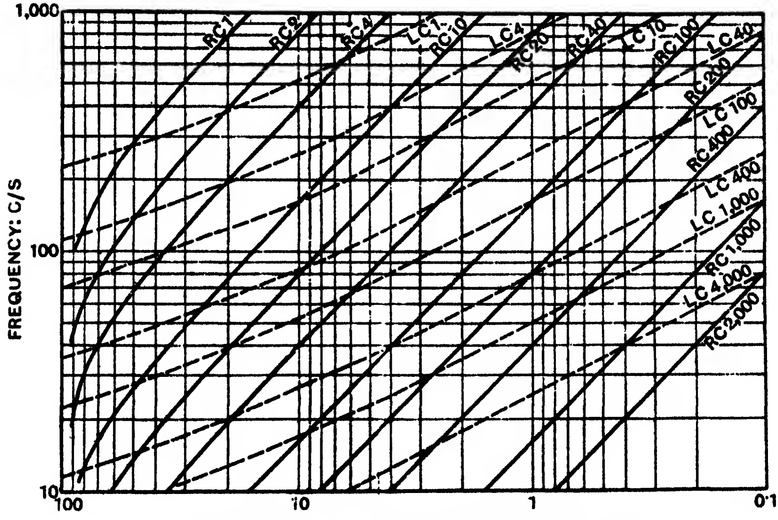


Fig. 14.29—Curves showing the effectiveness of the simple types of decoupler or hum filter (Fig. 14.28). The RC figures are C in μF multiplied by R in kΩ; LC figures are C in μF multiplied by L in H. It is assumed that the impedance of the load is much greater than that of C

In the *LC* filter,

$$\frac{V_{\text{in}}}{V_{\text{out}}} = \frac{X_L}{X_C} - 1$$

$$= 4\pi^2 f^2 LC - 1 \quad \text{(dotted curves in Fig. 14.29)}$$

Sectionalizing

If R/X_C exceeds 16 (fRC in kc/s–kΩ–μF > 2·5), it is advantageous to split R and C into equal ⅂ sections, cascaded. The advantage increases very rapidly with fRC:

Table 14.11

fRC (total) in kc/s–kΩ–μF	*Best No. of sections*	*Corresponding range of V_{in}/V_{out}*	*Range of V_{in}/V_{out} with RC all in one section*
Up to 2·5	1	Up to 16	Up to 16
2·5–7·2	2	16–130	16–45
7·2–14·3	3	130–1,000	45–90
14·3–23·6	4	1,000–7,500	90–150
23·6–35	5	7,500–55,000	150–223
35–50	6	55,000–450,000	223–310

For *LC* filters:

Table 14.12

f^2LC (*total*) *in* $(kc/s)^2$–*H*–μF	*Best No. of sections*	*Corresponding range of* V_{in}/V_{out}	*Range of* V_{in}/V_{out} *with LC all in one section*
Up to 0·6	1	Up to 22	Up to 22
0·6–1·7	2	22–250	22–66
1·7–3·4	3	250–2,800	66–135
3·4–6	4	2,800–34,000	135–230
6–9	5	34,000–450,000	230–360

Hum Voltage (Fig. 14.30)

With capacitance input filter, assuming current through rectifier into C_r flows as pulses at peaks of alternating voltage:

$$\text{Hum voltage} \simeq \frac{\sqrt{2}\,I}{(\omega^2LC-1)\omega C_r}$$

Provided that in every section of the filter $X_L/X_C >> 1$, the following rule is sufficiently accurate:
Divide $\sqrt{2}$ *times the output d.c.* (*in amps*) *by all the* ωL*'s*, ωC*'s and R's used for smoothing, including the reservoir* (C_r).

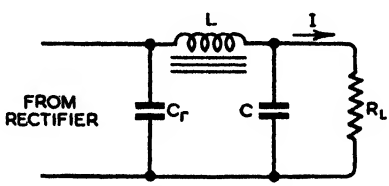


Fig. 14.30

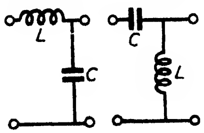
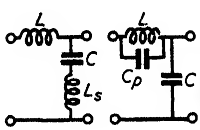
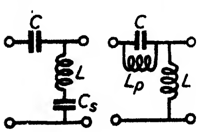
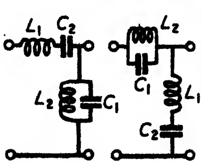
In this, ω can be taken as 2π times the fundamental hum frequency; i.e., 100 c/s in a 50-c/s full-wave rectifier. Note that *R* refers to the resistance of an *RC* section, not to choke-coil resistance, which is neglected.

Ref: " Smoothing Circuits ", *W.W.*, Oct. and Nov. 1949.

14.26. MATCHED-TERMINATION FILTERS

In Table 14.13 the formulae for component values for the chief varieties of filter network have been arranged more simply and concisely than usual by stating them all in terms of *half-section* arms. This introduces no difficulties in calculation, but should be kept in mind because elsewhere it is usual for similar symbols to refer to full-section

Table 14.13

Filter half-sections	*Component values*
	(R_0=input and output terminating voltage; ω_c= $2\pi\times$cut-off frequency; ω_∞= $2\pi\times$frequency of maximum attenuation; ω_1=$2\pi\times$lower cut-off frequency; ω_2=$2\pi\times$upper cut-off frequency)
Low-pass High-pass 	$L=\frac{R_0}{\omega_c}$ $C=\frac{1}{\omega_c R_0}$
Low-pass *m*-derived Series Parallel 	$m=\sqrt{1-\left(\frac{\omega_c}{\omega_\infty}\right)^2}$ $L=\frac{mR_0}{\omega_c}$ $C=\frac{m}{\omega_c R_0}$ $L_s=\frac{(1-m^2)R_0}{m\omega_c}$ $C_p=\frac{1-m^2}{m\omega_c R_0}$
High-pass *m*-derived Series Parallel 	$m=\sqrt{1-\left(\frac{\omega_\infty}{\omega_c}\right)^2}$ $L=\frac{R_0}{m\omega_c}$ $C=\frac{1}{m\omega_c R_0}$ $L_p=\frac{mR_0}{(1-m^2)\omega_c}$ $C_s=\frac{m}{(1-m^2)\omega_c R_0}$
Band-pass Band-stop 	$L_1=\frac{R_0}{\omega_2-\omega_1}$ $C_1=\frac{1}{(\omega_2-\omega_1)R_0}$ $L_2=\frac{(\omega_2-\omega_1)R_0}{\omega_1\omega_2}$ $C_2=\frac{\omega_2-\omega_1}{\omega_1\omega_2 R_0}$

arms. In either case, a factor of 2 or ½ has to be used for Π or T section arms.

For a half section (ᒣ or Γ), use Table values.

For a T section, use Table values for series (horizontal) arms, and half Table inductance and double Table capacitance for shunt (vertical) arms.

For a Π section, use Table values for shunt arms, and half Table capacitance and double Table inductance for series arms.

Theoretically, the same result can be obtained by either T or Π (Sec. 14.14), but one of them may have more practical component values. Parallel *m*-derived sections are likely to be better than series, especially at high frequencies, since coil self-capacitance can be merged in the parallel capacitance. The usefulness of *m*-derived sections is for heavily attenuating a frequency (f_∞) comparatively close to the cut-off frequency (f_c). Best impedance-matching is obtained with Π series and T parallel sections, but *m* must be the same for all sections, and preferably 0·6 or slightly over. T series and Π parallel sections of different *m* match one another but not resistances, unless half-sections are used at each end.

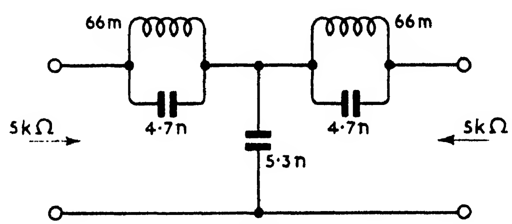


Fig. 14.31

E.g: Design a low-pass filter section to cut out 9-kc/s whistle and match 5 kΩ.

Since one particular frequency is to be rejected, the obvious choice is an *m*-derived type. As no cut-off frequency is mentioned, one is free to choose $m = 0{\cdot}6$, making a good match with 5 kΩ resistance:

$$0{\cdot}6 = \sqrt{1 - \left(\frac{f_c}{9}\right)^2}$$

$$\therefore f_c = 7{\cdot}2 \text{ kc/s}$$

$$L = \frac{mR_0}{\omega_c} = \frac{0{\cdot}6 \times 5}{2\pi \times 7{\cdot}2} = 0{\cdot}066 \text{ H}$$

$$C = \frac{m}{\omega_c R_0} = \frac{0{\cdot}6}{2\pi \times 7{\cdot}2 \times 5} = 0{\cdot}00266\ \mu\text{F}$$

$$C_p = \frac{1 - m^2}{m\omega_c R_0} = \frac{0{\cdot}64}{0{\cdot}6 \times 2\pi \times 7{\cdot}2 \times 5} = 0{\cdot}00472\ \mu\text{F}$$

Either Π series or T parallel would do, but the latter has the advantage of requiring much less coil winding. Following the rules for combining two half-sections gives the result shown (Fig. 14.31).

Refs: *Wave Filters*, by L. C. Jackson (Methuen).
Second Thoughts on Radio Theory, by "Cathode Ray" (Iliffe), Chap. 22.

14.27. TRANSMISSION LINES

Characteristic (or surge) impedance $Z_0 \simeq \sqrt{L/C}$ (losses neglected)

where L = inductance per unit length;

C = capacitance per unit length.

Parallel-wire Line (section, Fig. 14.32); air-spaced:

$L \simeq 0{\cdot}92 \log_{10}(2D/d)$ microhenries per metre

$C \simeq 12{\cdot}06/\log_{10}(2D/d)$ picofarads per metre

$Z_0 \simeq 276 \log_{10}(2D/d)$ ohms [$D \nless$ 4 or 5 times d]

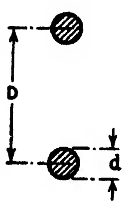


Fig. 14.32

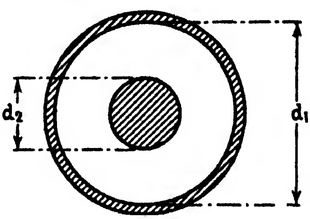


Fig. 14.33

Coaxial Line (section, Fig. 14.33); relative permittivity of spacing, ϵ_r

$L \simeq 0{\cdot}46 \log_{10}(d_1/d_2)$ microhenries per metre

$C \simeq 24{\cdot}1\ \epsilon_r/\log_{10}(d_1/d_2)$ picofarads per metre

$Z_0 \simeq (138/\sqrt{\epsilon_r}) \log_{10}(d_1/d_2)$ ohms

Velocity of propagation (phase velocity) $= c/\sqrt{\epsilon_r}$

c = velocity in free space

$\simeq 3 \times 10^8$ metres/sec

Wavelength along line $= \lambda/\sqrt{\epsilon_r}$

λ = wavelength in free space

Quarter-wave Transformer

A line of surge impedance Z_1 can be matched to another of Z_2 by linking them with a quarter-wave length of line of impedance $\sqrt{Z_1 Z_2}$.

Impedance of Loaded Line

The impedance Z_1 of a loss-free line of characteristic impedance Z_0 and length d (in wavelengths) terminated by an impedance Z_2 is

$$Z_1 = Z_0 \frac{Z_2 + jZ_0 \tan 2\pi d}{Z_0 + jZ_2 \tan 2\pi d}$$

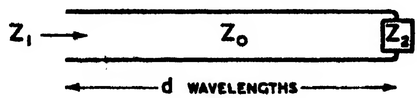


Fig. 14.34

Refs: *High Frequency Transmission Lines*, by Willis Jackson (Methuen).
Second Thoughts on Radio Theory, by " Cathode Ray " (Iliffe), Chap. 23.

14.28. AERIALS

Effective height of rectangular frame aerial, in metres

$$= 2NH \sin\left(\frac{\pi f W}{300{,}000}\right)$$

N = number of turns.
H = height in metres.
W = width in metres.
f = frequency in kc/s.

When multiplied by the field strength in μV/m the above gives the microvoltage induced in the frame aerial (see Sec. 11.28).

Formulæ for open aerials are not given, because only ideal forms can be calculated with reasonable accuracy.

Effective height of ferrite rod aerial, in metres

$$= \frac{2\pi f N A F \mu}{300{,}000}$$

f = frequency in kc/s
N = number of turns
A = area of each turn in sq. metres
F = ratio of mean to maximum flux density in coil
μ = effective permeability of rod (relative to 1 for air)

14.29. CATHODE RAY DEFLECTION

Velocity of electron = $594\sqrt{V}$ km/sec
[$V \not> 10{,}000$, beyond which increase in electron mass due to velocity is appreciable]

Electric Deflection

$$D = \frac{lLv}{2Vd}$$

[D, l, L, d all in same units]

l = length of deflecting field
L = length from centre of field to screen
v = difference of potential between deflection plates
d = distance between plates
V = accelerating voltage
H = flux density

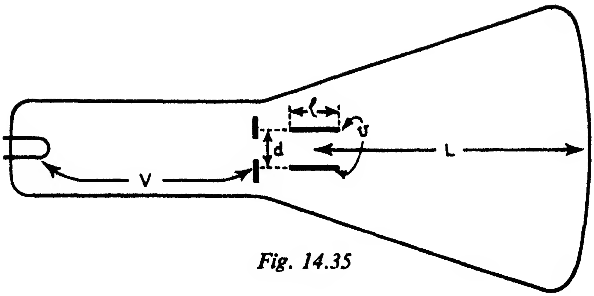


Fig. 14.35

Magnetic Deflection

$$D \simeq \frac{0{\cdot}3lLH}{\sqrt{V}}$$ [D, l, L, all in cm; H in gauss. If H in Wb/m^2, multiply by 10^4]

14.30. LIGHT

Velocity

$c = 299{,}792 \pm 2$ km/s (International Standard).

Ref: *W.W.*, Dec. 1954, p. 590.

Visible Wavelength Range (Fig. 14.36)

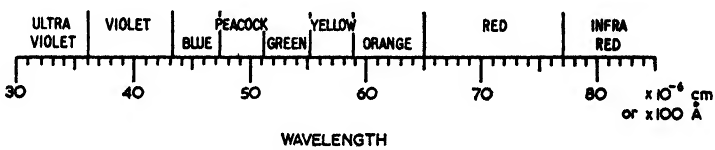


Fig. 14.36

Units

Light units are based on the idea of *luminous flux* streaming equally in all directions from a point source. The amount of flux is proportional to the *luminous intensity* of the source. The unit of luminous intensity is the *candela*, equal to that from 1/60 cm^2 of platinum at its melting temperature. It is 1·9% smaller that its predecessor, the *international candle*. The total flux emitted by a source of 1 candela is 4π *lumens*, so 1 lumen is the flux flowing through unit area at unit radius from unit source. The unit of flux density received at right angles by a surface—its *illumination*—equal to 1 lumen/m^2 is the *lux*. The foot-candle, still sometimes used, is approximately an illumination of 1 lumen/ft^2 (= 10·76 lux). *Luminance* is the brightness of a surface, or its luminous intensity per unit area (as seen from the receiving point). The brightness of a totally-reflecting matt surface illuminated by 1 lux is 1 apostilb. Also used are the *lambert* (= 10^4 apostilbs, or 1 lumen/cm^2) and the foot-lambert (= 1·076 millilambert or 1 lumen/ft²). *Candle-power* is now the light-radiating capacity of a source in a given direction, in terms of the luminous intensity in candelas.

Ref: B.S. 233 : 1953. *Glossary of Terms used in Illumination and Photometry* (B.S.I.).

14.31. SOUND

Velocity

In air, $1{,}090 + 2\,T_0$ ft/sec $\qquad T_0$ = temp in °C

Intensity

The rate of flow of sound energy per unit area at right angles to direction of propagation. (Electrical circuit analogue: power.) M.k.s. unit: watt/m². 1 watt/m² = 1,000 c.g.s. units (ergs/sec/cm²).

10^{-12} watts/m² is called the threshold intensity, being the intensity of a 1-kc/s sound that can just be heard by a person with normal hearing; and dB ratings of intensity are reckoned with reference to it. In air at 20°C and 0·76 metre of mercury, the r.m.s. sound pressure (electrical circuit analogue: voltage) corresponding to threshold intensity is 204×10^{-6} dyne/cm² or $20{\cdot}4 \times 10^{-6}$ newton/m². (The dyne/cm² is sometimes called the bar, but this is confusing, because the internationally-adopted bar is 10^6 times as great.)

Table 14.14

dB above threshold	*Intensity*			*Corresponding pressure in air**	
	W/m²	*μW/cm²*	*ergs/sec/cm²*	*dynes/cm²*	*newton/m²*
120	1	100	1000	200	20·0
100	10^{-2}	1	100	20·0	2·0
80	10^{-4}	10^{-2}	0·1	2·0	0·2
60	10^{-6}	10^{-4}	10^{-3}	0·2	0·02
40	10^{-8}	10^{-6}	10^{-5}	0·02	0·002
20	10^{-10}	10^{-8}	10^{-7}	0·002	0·0002
0	10^{-12}	10^{-10}	10^{-9}	0·0002	0·00002

* The pressure corresponding to a given intensity depends on the medium, and in air at 20°C and 0·76 m of mercury is 2% higher than the figures shown, which nevertheless are the usual standard.

The sensitivity of microphones is usually specified as dB below 1 volt per dyne/cm².

Loudness

Loudness *level*, or equivalent loudness, is numerically equal to the intensity in dB of a 1-kc/s sound judged by a person of normal hearing to be equally loud. The loudness level scale therefore coincides with the intensity scale at 1 kc/s, but, owing to the characteristics of the human ear, differs at other frequencies, as shown in the Fletcher-

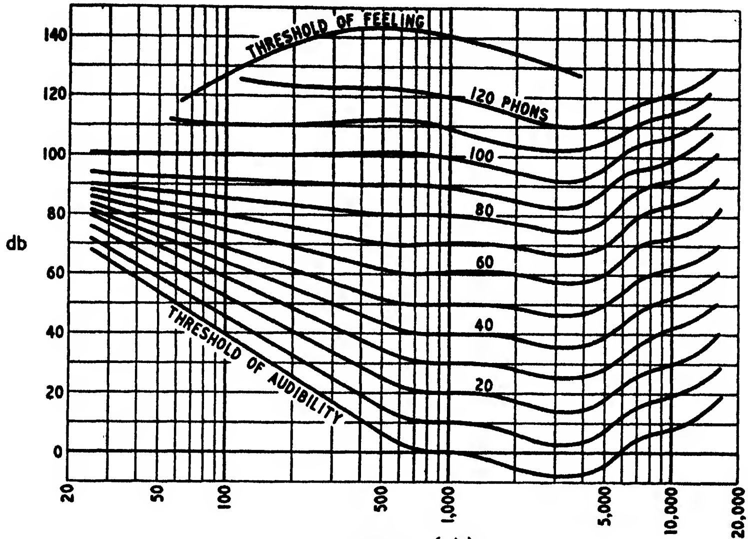


Fig. 14.37—Fletcher-Munson equal-loudness curves, obtained by tests on listeners with normal hearing. The number of phons specifies the loudness of a pure 1 kc/s sound having an intensity of the same number of dB above the threshold of audibility

Munson curves (Fig. 14.37).* Sometimes, especially in America, loudness level is specified in dB, but since the figures agree with intensity dB only at 1 kc/s, this practice is most confusing, and use of the name *phon* for the unit of loudness level is preferable.

Although the loudness level has a subjective basis, it is tied to objective intensity at 1 kc/s and its figures do not accurately indicate the relative loudnesses of sounds. *Loudness* is therefore specified in *sones*, which are related to phons by this empirical formula:

$$\log_{10}S \simeq 0{\cdot}029\ (P - 40)$$

where S is the number of sones and P the number of phons.

Refs: B.S. 661 : 1955, *Glossary of Acoustical Terms*;
B.S. 2497 : 1954, *The Normal Threshold of Hearing for Pure Tones by Earphone Listening*;
B.S. 3045 : 1958 *The Relation between the Sone Scale of Loudness and the Phon Scale of Loudness Level* (B.S.I.).
"Loudness", *W.W.*, Nov. 1957.

14.32. NOISE

Thermal-agitation (Johnson) Noise

$$E^2 = 4kTBR$$

where E = r.m.s. value of noise e.m.f. in circuit of resistance R at

* Greater accuracy is claimed for curves by Robinson and Dadson, "The Measurement of Loudness", *Jour. Acoust. Soc. Amer.*, Sept. 1955.

temperature T °K, within an effective bandwidth B c/s;

and k = Boltzmann's constant = $1{\cdot}38 \times 10^{-23}$ joules/°K.

Noise power is distributed uniformly over all frequencies.

" Available noise power ", the power that would be delivered to a matched load

$$R_L\ (=R) = E^2/4R = kTB.$$

T, if room temperature, is usually taken as 290 (= 63° F).

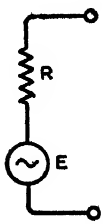


Fig. 14.38—The fluctuation noise produced by a resistance R is represented by an imaginary generator producing the same r.m.s. e.m.f. E

Shot (Valve) Noise

$$I^2 = 2\,e\,I_a\,B$$

where I = r.m.s. value of noise current in anode current I_a of temperature-limited diode;

e = charge on electron

= $1{\cdot}6 \times 10^{-19}$ coulombs;

and B = effective bandwidth, in c/s.

The space charge in a valve used under usual conditions considerably reduces noise. In tetrodes and other multi-electrode valves there is additional noise due to random partition of the space current.

Noise Factor, N, of a Receiver or Amplifier

$$N = \frac{\text{available signal/noise ratio at input}}{\text{signal/noise ratio at output}}$$

$$= \frac{\text{total noise output}}{\text{noise output due to source (e.g., aerial)}}$$

All quantities reckoned as *power*. If there were no Johnson or shot noise generated in the amplifier, N would = 1.

Refs: " Noise ", *W.W.*, May and June 1952.
" Valve Noise ", *W.W.*, Dec. 1960.
" Transistor Noise ", *W.W.*, Jan. 1961.

14.33. MATHEMATICAL FORMULAE*

Circles and Spheres

r = radius; d = diameter; $\pi = 3{\cdot}1416 \ldots$

$$\text{Circumference of circle} = 2\pi r = \pi d$$

$$\text{Area of circle} = \pi r^2 = \frac{\pi d^2}{4}$$

$$\text{Surface area of sphere} = 4\pi r^2 = \pi d^2$$

$$\text{Volume of sphere} = \frac{4\pi r^3}{3} = \frac{\pi d^3}{6}$$

Angles

$$1 \text{ radian} = \frac{180^\circ}{\pi} = 57{\cdot}3^\circ$$

In mathematical formulae, angles are in radians.

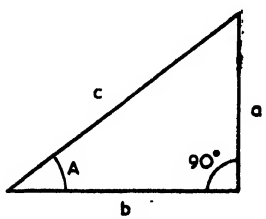


Fig. 14.39

$$\sin A = \frac{a}{c} \qquad \operatorname{cosec} A = \frac{c}{a}$$

$$\cos A = \frac{b}{c} \qquad \sec A = \frac{c}{b}$$

$$\tan A = \frac{a}{b} \qquad \cot A = \frac{b}{a}$$

Table 14.15

Angle A {*degrees* / *radians*}	0 0	30 $\pi/6$	45 $\pi/4$	60 $\pi/3$	90 $\pi/2$	180 π	270 $3\pi/2$	360 2π
sin *A*	0	$\frac{1}{2}$	$\sqrt{2}/2$	$\sqrt{3}/2$	1	0	-1	0
cos *A*	1	$\sqrt{3}/2$	$\sqrt{2}/2$	$\frac{1}{2}$	0	-1	0	1
tan *A*	0	$\sqrt{3}/3$	1	$\sqrt{3}$	$\pm\infty$	0	$\pm\infty$	0

* For symbols, see p. 448.

$$\sin(A \pm B) = \sin A \cos B \pm \cos A \sin A$$

$$\cos(A \pm B) = \cos A \cos B \mp \sin A \sin A$$

$$\sin A + \sin B = 2 \sin \tfrac{1}{2}(A+B) \cos \tfrac{1}{2}(A-B)$$

$$\cos A + \cos B = 2 \cos \tfrac{1}{2}(A+B) \cos \tfrac{1}{2}(A-B)$$

$$\sin A \sin B = \tfrac{1}{2}[\cos(A-B) - \cos(A+B)]$$

$$\cos A \cos B = \tfrac{1}{2}[\cos(A+B) + \cos(A-B)]$$

$$\sin A \cos B = \tfrac{1}{2}[\sin(A+B) + \sin(A-B)]$$

$$\sin^2 A + \cos^2 A = 1$$

$$\sin^2 A = \tfrac{1}{2}(1 - \cos 2A)$$

$$\cos^2 A = \tfrac{1}{2}(1 + \cos 2A)$$

$$\sin 2A = 2 \sin A \cos A$$

$$\cos 2A = \cos^2 A - \sin^2 A$$

(These formulae are familiar in connection with modulation and sidebands.)

Quadratic Equation

$$ax^2 + bx + c = 0$$

Solution:
$$x = \frac{-b \pm \sqrt{b^2 - 4ac}}{2a}$$

Logarithms and Exponentials

$y = n^x$ ("n to power x") and $x = \log_n y$ ("log y to base n") are alternative ways of saying the same thing. A logarithm is an index or exponent; its usefulness lies in the fact that adding indices multiplies the main terms; e.g., $n^a \times n^b = n^{a+b}$. For calculations, tables and slide rules embody " common " logarithms (base 10) for practical convenience; in mathematical relationships, logarithms are " natural " (base e).*

$$e = 1 + 1 + \frac{1}{2!} + \frac{1}{3!} + \frac{1}{4!} \ldots\ldots = 2{\cdot}71828 \ldots$$

It is easy to transfer from one base to another, for $\log_m y = \log_n y \log_m n$

e.g: $\log_e y = \log_{10} y \log_e 10 = \log_{10} y \times 2{\cdot}3026.$

Conversely, $\log_{10} y = \log_e y \times 0{\cdot}4343.$ (See Fig. 14.40*a*.)

* When no base is specified it must be inferred from the context. In mathematical theory it is likely to be e, but there is a tendency to use " log x " when the base is 10 and to write the logarithm to base e as " ln. x ".

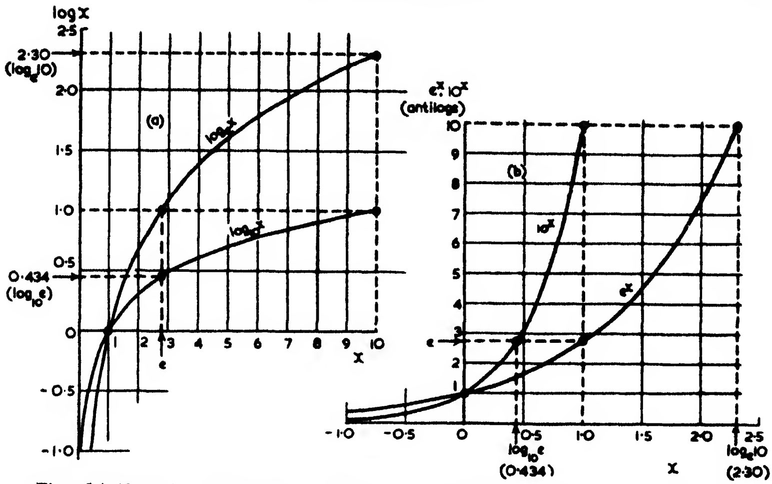


Fig. 14.40—(a) *Logarithmic and* (b) *exponential curves, for 10 and* e. (b) *is the same as* (a) *turned over a diagonal line*

The following *exponential functions* (Fig. 14.40*b*) are much used in electrical theory:

$$e^{x} = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \frac{x^4}{4!} + \ldots$$

$$e^{jx} = \cos x + j \sin x$$

(represents a vector of unit length, at angle x radians)

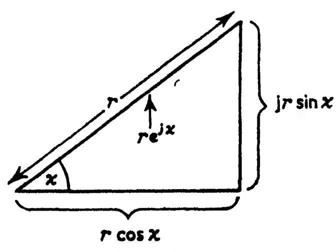


Fig. 14.41

$$e^{-jx} = \cos x - j \sin x$$

$$\sin x = \frac{e^{jx} - e^{-jx}}{2j}$$

$$\cos x = \frac{e^{jx} + e^{-jx}}{2}$$

Ref. "e", *W. W.*, April 1960.

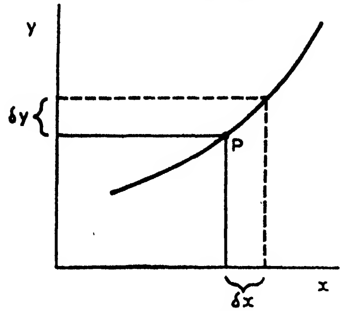


Fig. 14.42

Calculus

The slope of a graph is a measure of the rate at which one variable (y) is changing relative to another (x). At any point P the slope $\simeq \delta y/\delta x$, provided that the curve is continuous and δy and δx are sufficiently

Table 14.16

y	$\dfrac{dy}{dx}$	$\int y.dx$	$\dfrac{x\,dy}{y\,dx}$
a	0	ax	0
x^n	nx^{n-1}	$\dfrac{x^{n+1}}{n+1}$ $\quad n\neq -1$	n
ax^n+b	nax^{n-1}	$\dfrac{ax^{n+1}}{n+1}+bx$ $\quad n\neq -1$	$\dfrac{n}{1+b/ax^n}$
e.g: ax $\quad n=1$	a	$ax^2/2$	1
ax^2 $\quad n=2$	$2\,ax$	$ax^3/3$	2
ax^3 $\quad n=3$	$3\,ax^2$	$ax^4/4$	3
$a\sqrt{x}$ $\quad n=\frac{1}{2}$	$a/2\sqrt{x}$	$\frac{2}{3}\,ax\sqrt{x}$	$\frac{1}{2}$
a/x $\quad n=-1$	$-a/x^2$	$a\log_e x$	-1
a/x^2 $\quad n=-2$	$-2a/x^3$	$-a/x$	-2
$a/\sqrt{x}$ $\quad n=-\frac{1}{2}$	$-a/2x\sqrt{x}$	$2a\sqrt{x}$	$-\frac{1}{2}$
$\sin(ax+b)$	$a\cos(ax+b)$	$-\dfrac{1}{a}\cos(ax+b)$	$ax/\tan(ax+b)$
$\cos(ax+b)$	$-a\sin(ax+b)$	$\dfrac{1}{a}\sin(ax+b)$	$-ax\tan(ax+b)$
$\log_e(ax+b)$	$\dfrac{a}{ax+b}$	$\left(x+\dfrac{b}{a}\right)\log_e(ax+b)-x$	$\dfrac{ax}{(ax+b)\log_e(ax+b)}$
e^{ax}	ae^{ax}	e^{ax}/a	ax
$u+v$	$\dfrac{du}{dx}+\dfrac{dv}{dx}$	$\int u.dx+\int v.dx$	
uv	$u\dfrac{dv}{dx}+v\dfrac{du}{dx}$		
$\dfrac{u}{v}$	$\left(v\dfrac{du}{dx}-u\dfrac{dv}{dx}\right)\Big/v^2$		
$f(u)$	$\dfrac{dy}{du}\cdot\dfrac{du}{dx}$		

E.g.: Differentiate $\dfrac{x^2}{x+1}$

This is covered by the procedure (on the left)

for $\dfrac{u}{v}$: $\quad \dfrac{(x+1)\times 2x-(x^2\times 1)}{(x+1)^2}$

$$=\frac{x(x+2)}{(x+1)^2}$$

small. When they are made infinitesimally small they are denoted by dy and dx; and dy/dx is the exact slope, or differential coefficient, or derivative, of y with respect to x. $\frac{d}{dx}$ is an operator, alternatively denoted by D_x. Another notation is, for the original equation, $y = f(x)$ (i.e., y = a function of x) and, for the derivative, $dy/dx = f'(x)$.

The reverse process, symbolized by $\int dx$, is integration. $\int y.dx$ is that function of x, the slope of whose curve is y. Since the graph of a constant term has zero slope, the value of such a term (if any) is not revealed by integration, and can be found only from additional data. In Table 14.16 it is omitted, but should not be forgotten. a and b denote constants; and y, u, and v are functions of x. Note that a constant factor in the original function appears as such in the derivative and the integral. The use of the last column is explained below under "Errors and Approximations".

Maxima and Minima

Where a function has a maximum or minimum value its slope is zero; so, to find such values, differentiate the function and equate the result to zero. This is a most valuable use of differentiation.

E.g: The power in a load R_L, from a generator having an e.m.f. E and resistance R, is $E^2R_L/(R_L+R)^2$; for what value of R_L is the power in it a maximum? The u/v formula above applies, and is zero when

$$v\frac{du}{dx}=u\frac{dv}{dx};$$

$$\text{i.e.}\quad (R_L+R)^2\,E^2=E^2R_L(2R_L+2R),$$

which simplifies to $R_L=R$, the condition for maximum power. (Common sense indicates that it is not a minimum.)

Errors and Approximations

It is often useful to know how much a small variation or error in one value will affect another. Suppose δy is the change in the value of y caused by a change δx in the value of x. Then if δx is small enough for the slope of the x, y, curve to be practically constant over that range,

$$\delta y/\delta x \simeq dy/dx,$$

so
$$\delta y \simeq \delta x \times dy/dx$$

The proportionate change, $\frac{\delta y}{y}$, is therefore $\simeq \frac{\delta x}{x}.\frac{x}{y}.\frac{dy}{dx}$. The last column in Table 14.16 gives $\frac{x}{y}.\frac{dy}{dx}$ for various functions, and represents the approximate ratio of $\delta y/y$ to $\delta x/x$, or the percentage change in y for 1% change in x.

E.g: What is the effect on frequency of a small increase in oscillator tuning capacitance?

$f_0 = 1/(2\pi\sqrt{LC})$, which has the form "$y = a/\sqrt{x}$", for which Table 14.16 shows $\frac{x}{y}\cdot\frac{dy}{dx} = -\frac{1}{2}$, meaning that for every 1% increase in C, f_0 *de*creases approximately $\frac{1}{2}$%. The actual change in frequency, δf_0, $\simeq \delta C\frac{df_0}{dC} = -\frac{\delta C}{2\pi\sqrt{LC}\times 2C}$

Although this is an approximation, it is generally more accurate than calculating y and $y+\delta y$ and taking the difference.

Useful approximations when $x << 1$:

$$\sin x \simeq x \qquad \cos x \simeq 1 \qquad \tan x \simeq x$$

$$e^x \simeq 1+x \quad \text{or} \quad \log_e(1+x) \simeq x$$

$$e^{-x} \simeq 1-x \quad \text{or} \quad \log_e(1-x) \simeq -x$$

Hyperbolic Functions

$$\sinh x = \frac{e^x - e^{-x}}{2} \qquad \cosh x = \frac{e^x + e^{-x}}{2} \qquad \tanh x = \frac{\sinh x}{\cosh x}$$

$$\sinh jx = j\sin x \qquad \cosh jx = \cos x \qquad \tanh jx = \frac{j\sin x}{\cos x}$$

$$\sinh(\alpha + j\beta) = \sinh\alpha\cos\beta + j\cosh\alpha\sin\beta$$

$$\cosh(\alpha + j\beta) = \cosh\alpha\cos\beta + j\sinh\alpha\sin\beta$$

These are commonly used in transmission-line calculations.

Ref.: "h", *W. W.*, May 1960.

14.34. DECIBELS (AND NEPERS)

One way of expressing the relative strengths of two different signals, or the same signal at different times, is to say that one of them is x times greater than the other. A disadvantage of this method of reckoning is that if the relationship between gain and, say, frequency is expressed by means of a curve, it is very difficult to get a fair idea of the performance of the amplifier by looking at the shape of the curve. Fig. 14.43 shows three frequency characteristic curves. The only thing that can be reliably learnt from a glance at them is that amplifier A gives a greater gain than B, which in turn gives more than C. But whereas it appears at first sight that the difference between A and B is wider than that between B and C, reference to the scale of gain shows that while A is $2\frac{1}{2}$ gimes B, B is 4 times C. This way of plotting amplification curves is therefore misleading, even if one is interested in comparing merely the general level of gain. But in considering a frequency characteristic the actual gain is of less consequence than the relative gain at different frequencies. Curve C is nearly flat compared with curve A, and one might jump to the conclusion that the amplifier it

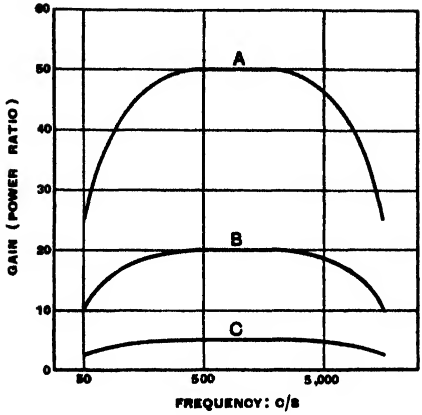


Fig. 14.43—Three characteristic curves plotted on an ordinary magnification or gain scale. The general appearance of the curves leads to quite wrong conclusions

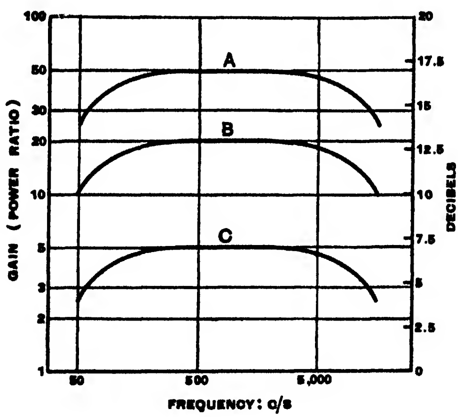


Fig. 14.44—When the curves of Fig. 14.43 are replotted on a logarithmic gain scale, which is the same as a linear decibel scale, the true proportions appear

represents gives more uniform amplification. But in actual fact all three curves indicate identical frequency characteristics.

As a plotted curve has little purpose if it does not succeed in saving one from having to make numerical calculations in order to draw fair conclusions, there is a need for a system of reckoning gain which will make things that actually are equal look equal. This object is attained if gain is plotted on logarithmically-divided paper. Fig. 14.44 shows the same three curves transferred to such a sheet, and even if the gain scale were omitted altogether one could depend upon the shapes and respective levels of the three curves to present a fair comparison. A given distance along the vertical scale represents a certain *increase* in gain, wherever along the scale it may be; so it is an obvious step to divide the scale into equal divisions and call them units of gain.

A tenfold power increase is divided into ten equal steps called decibels, and as a gain of 1 on the original scale is neither a gain nor a loss, it is 0 on the decibel scale. 10 on the gain ratio scale therefore corresponds to 10 on the dB scale; 100 corresponds to 20, and so on. Because the dB scale is uniformly divided, one is free to put the zero wherever it is convenient. For instance, in studying one of the characteristics shown in Fig. 14.44 it might be convenient to fix the zero level by the flat portion of the curve, or by its level at some standard frequency such as 400 or 1,000 c/s. The graph would then directly read dB below or above normal. And for curves of such things as gramophone pick-ups, where gain is meaningless, the choice of zero dB level is similarly open.

For this reason it is nonsense to refer to a power or signal strength of so many decibels, unless some starting-point is specified or understood. It is correct to refer to an output of half a watt as + 10 dB *if it has been agreed to call* 50 *mW* " *zero* ". The abbreviation " dB(mW) " (or " dBm ") is used to mean decibels with reference to 1mW.

Mathematically, the gain in dB $= 10 \log_{10}(P_2/P_1)$, where P_1 and P_2 are the two powers being compared. In radio practice it is usual to measure the signal at one or both ends of the amplifier in *volts*. In line-telephone work, where decibels originated, signals are usually established across some definite impedance such as 600 Ω resistance, and as power is proportional to V^2, the gain in dB $= 20 \log_{10} (V_2/V_1)$, *assuming that the impedance is the same in both cases*. If V_1 and V_2 are before-and-after figures, being measured at the same place each time, there is obviously no difficulty about impedance. But if the impedances are different the proper formula is $20 \log_{10} (V_2/V_1) + 10 \log_{10} (R_1/R_2)$, where R_1 and R_2 are the resistive components of the impedances; or if currents I_1 and I_2 are measured, the gain in dB is $20 \log_{10} (I_2/I_1) + 10 \log_{10} (R_2/R_1)$; otherwise one gets absurdities such as a high-gain amplifier with a low impedance output apparently giving a decibel loss, or a transformer by itself giving a gain.

In valve amplifiers the impedance at the input to a valve may sometimes be practically infinity, which leads to a result almost as absurd as the previous one, and even less useful; so, as the actual gain or loss at any particular point on the curve is generally of minor importance compared with the relative level under various conditions, it is customary to leave impedance out of account. But this practice must not be allowed to obscure the true principle, or confusion sooner or later is certain.

Table 14.17 connects decibels with power and voltage (or current) ratios, but the best way is to read them off a slide-rule. Suppose the

Table 14.17

The decibel figures are in the centre column: figures to the left represent decibel loss, and those to the right decibel gain. The voltage and current figures are given on the assumption that there is no difference in impedance.

Voltage or current ratio	*Power ratio*	← − *dB* + →	*Power ratio*	*Voltage or current ratio*
1·000	1·000	0	1·000	1·000
0·989	0·977	0·1	1·023	1·012
0·977	0·955	0·2	1·047	1·023
0·966	0·933	0·3	1·072	1·035
0·955	0·912	0·4	1·096	1·047
0·944	0·891	0·5	1·122	1·059
0·933	0·871	0·6	1·148	1·072
0·912	0·832	0·8	1·202	1·096
0·891	0·794	1·0	1·259	1·122
0·841	0·708	1·5	1·413	1·189
0·794	0·631	2·0	1·585	1·259
0·750	0·562	2·5	1·778	1·334
0·708	0·501	3·0	1·995	1·413
0·668	0·447	3·5	2·239	1·496
0·631	0·398	4·0	2·512	1·585
0·596	0·355	4·5	2·818	1·679
0·562	0·316	5·0	3·162	1·778
0·501	0·251	6·0	3·981	1·995
0·447	0·200	7·0	5·012	2·239
0·398	0·159	8·0	6·310	2·512
0·355	0·126	9·0	7·943	2·818
0·316	0·100	10	10·00	3·162
0·282	0·0794	11	12·6	3·55
0·251	0·0631	12	15·9	3·98
0·224	0·0501	13	20·0	4·47
0·200	0·0398	14	25·1	5·01
0·178	0·0316	15	31·6	5·62
0·159	0·0251	16	39·8	6·31
0·126	0·0159	18	63·1	7·94
0·100	0·0100	20	100·0	10·00
$3{\cdot}16\times10^{-2}$	10^{-3}	30	10^{3}	$3{\cdot}16\times10$
10^{-2}	10^{-4}	40	10^{4}	10^{2}
$3{\cdot}16\times10^{-3}$	10^{-5}	50	10^{5}	$3{\cdot}16\times10^{2}$
10^{-3}	10^{-6}	60	10^{6}	10^{3}
$3{\cdot}16\times10^{-4}$	10^{-7}	70	10^{7}	$3{\cdot}16\times10^{3}$
10^{-4}	10^{-8}	80	10^{8}	10^{4}
$3{\cdot}16\times10^{-5}$	10^{-9}	90	10^{9}	$3{\cdot}16\times10^{4}$
10^{-5}	10^{-10}	100	10^{10}	10^{5}
$3{\cdot}16\times10^{-6}$	10^{-11}	110	10^{11}	$3{\cdot}16\times$
10^{-6}	10^{-12}	120	10^{12}	10^{6}

output power of a receiver is observed to rise from 46 to 58 mW as the result of a change in frequency. The ratio found on the slide-rule in the ordinary way is 1·26; but if instead of this the log scale is read—nearly 0·1—the gain in dB is obtained by the simple process of multiplying by 10, and is therefore 1. This works only when the ratio is not over 10. For every additional figure to the left of the decimal point in the ratio it is necessary to add 10 dB to the result; or to subtract 10 for each place to the right. With practice the process becomes as rapid as ordinary slide-rule calculation. If the readings in the above example were taken in volts they would be in the ratio of 1·12 (i.e., $\sqrt{1{\cdot}26}$), or 0·05 on the log scale, and must be multiplied by 20, giving 1 dB again. Incidentally, 1 dB is about the least change in signal strength that can be noticed by ear when it is made rapidly.

Nepers

Decibels are power ratios expressed as common logarithms. In fundamental line calculations it is more convenient to work in nepers, which are current ratios expressed as natural logarithms:

$$\text{Number of nepers} = \log_e (I_2/I_1) \quad \text{or} \quad \tfrac{1}{2}\log_e (P_2/P_1)$$

Given equal impedance:

$$1\ \text{N} = 8{\cdot}686\ \text{dB} \qquad 1\ \text{dN (decineper)} = 0{\cdot}8686\ \text{dB}$$
$$1\ \text{dB} = 0{\cdot}1151\ \text{N} \qquad 1\ \text{dB} = 1{\cdot}151\ \text{dN}$$

14.35. MUSICAL INTERVALS AND FREQUENCIES

The natural scale for frequency of sound is also logarithmic. It is interesting to note that a whole tone in music is practically the same ratio of frequency as the voltage ratio equal to 1 dB; a semitone corresponds to half a dB. Table 14.18 gives the frequency ratios

Table 14.18

Note		*Interval from lowest*	*Fractional Ratios*	*Frequency Ratio*
C^1	Doh	Octave	48 : 24, or 2	2·000
B	Te	Seventh	45 " " $\frac{15}{8}$	1·888
A	Lah	Sixth	40 " " $\frac{5}{3}$	1·682
G	Soh	Fifth	36 " " $\frac{3}{2}$	1·498
F	Fah	Fourth	32 " " $\frac{4}{3}$	1·335
E	Me	Third	30 " " $\frac{5}{4}$	1·260
D	Ray	Second	27 " " $\frac{9}{8}$	1·122
C	Doh	Unison	24 " " 1	1·000
		Tone	27 " " $\frac{9}{8}$	1·122
		Semitone	51 : 48, " $\frac{17}{16}$	1·059

corresponding to all the intervals in one octave of the scale of C major, which is played on the white keys of the piano. The actual frequencies of the notes are shown in Fig. 14.45 (overleaf). The frequency ratios are calculated on the equal-temperament scale, to which keyboard instruments are tuned to allow changes of key without retuning; as can be seen, they do not agree *exactly* with the fractional ratios, which are correct as judged by ear.

Ref.: B.S. 661 : 1955, *Glossary of Acoustical Terms*, Sec. 9 (B.S.I.).

14.36. FREQUENCY ALLOCATIONS

International Allocations from May 1961

Those shared with other services are marked S. All are for world-wide use except those marked R, which are effective in Region 1 (Europe, U.S.S.R. and Africa). Region 2 is N. and S. America, and Region 3 Asia (except U.S.S.R.) and Australasia. There are some special provisions for particular countries, such as shared use of v.h.f. channels.

Table 14.19

Broadcasting		*Amateur*	*Standard Frequency*
SR 150–160 kc/s	11700–11975 kc/s	S 3500–3800 kc/s	2498–2502 kc/s
R 160–255	15100–15450	7000–7100	4995–5005
SR 255–285	17700–17900	S 7100–7150	9995–10005†
R 525–535	21450–21750	14000–14350	14990–15010
535–1605	25600–26100	21000–21450	19990–20010†
S 2300–2495	SR 41–47 (I)	28000–29700	24990–25010
SR 2495–2498	R 47–68 Mc/s (I)	144–146 Mc/s	
S 3200–3400	87·5–88 (II)	S 420–440	†Includes allocations on secondary basis to space research.
	88–100* (II)		
SR 3950–4000	SR 174–223* (III)	SR 450–460	
SR 4750–4850	470–582 (IV)	S 1215–1300	
S 4850–4995	SR 582–606	S 2300–2450	
S 5005–5060	R 606–790 (V)	S 5650–5850	
5950–6200	S 790–890 (V)	S10000–10500	
R 7100–7300	SR 890–960 (V)	21–22 Gc/s	
9500–9775	S 11·7–12·7 Gc/s	Also, in certain countries including U.K., up to 200 kc/s in the band 1,715–2,000 kc/s; max. power 10 W.	

Note: v.h.f. broadcasting band numbers are shown in brackets.

* 95-100 and 216–223 Mc/s not available for broadcasting in U.K.

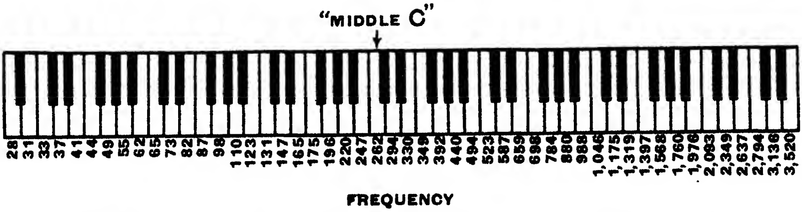


Fig. 14.45—Piano scale showing the frequencies to which the keys are tuned. The usually published scales based on C = 256 are misleading. Frequencies of black keys can be obtained by multiplying the frequency of the next white key below by 1.059

British Television Channels

Table 14.20.

Television Sound Channel Frequencies

Band	*Channel*	*Sound Mc/s*	*Vision Mc/s*
I	1	41·50	45·00
	2	48·25	51·75
	3	53·25	56·75
	4	58·25	61·75
	5	63·25	66·75
III	6	176·25	179·75
	7	181·25	184·75
	8	186·25	189·75
	9	191·25	194·75
	10	196·25	199·75
	11	201·25	204·75
	12	206·25	209·75
	13	211·25	214·75

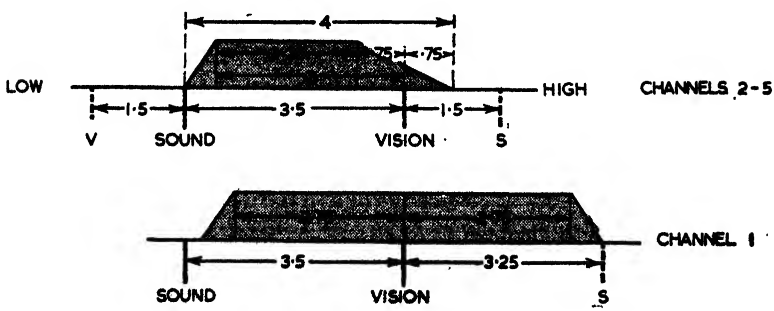


Fig. 14.46

14.37. STANDARD FREQUENCIES

B.B.C. Transmitters

The Droitwich transmitter of the British Broadcasting Corporation is guaranteed to maintain its nominal frequency of 200 kc/s within 1 in 10^7.

All B.B.C. medium-frequency transmitters are usually within 1 in 10^7, although this is not guaranteed.

The tuning note preceding regular Home Service and Light Programme transmissions is 1,000 c/s $\pm$ 1 in 10^8. The Third Programme transmissions are preceded first (at 17.45 hours) by 1,000 c/s and then by 440 c/s, both $\pm$ 1 in 10^8.

U.K. Standard-frequency Transmissions

These signals are radiated daily from G.P.O. transmitters at Rugby, call sign MSF, to the following schedule:

Frequency	*Power*	*Time of Transmission*
60 kc/s	10 kW	14.29—15.30 G.M.T.
2·5 Mc/s 5 Mc/s 10 Mc/s	0·5 kW	Continuous except between 15 and 20 minutes past each hour

Each transmission is divided up as follows:

Minutes past each hour				*Modulation*
0–5		30–35	45–50	1,000 c/s
5–10	20–25	35–40	50–55	1 c/s, 59th omitted
10–14	25–29	40–44	55–59	Unmodulated
14–15	29–30	44–45	59–60	Announcement

These transmissions (and also those of the G.P.O. time-signal transmitter at Rugby, GBR, on 16 kc/s,* and associated short-wave stations operating to schedules issued every few months) are controlled by the same standard oscillator, calibrated to an accuracy of $\pm$ 2 in 10^{10} against the caesium atomic resonance frequency, taken as 9,192,631,770 $\pm$ 20 c/s. For most purposes the radiated frequencies are sufficiently accurate, but corrections to $\pm$ 1 in 10^{10} are published monthly in *Electronic Technology*. Full information on the MSF service is obtainable from The Director, National Physical Laboratory, Teddington, Middlesex. See also *E.T.*, March 1959, p. 117.

* Criggion (GBZ) radiates on 19·6 kc/s when GBR is inoperative.

American Bureau of Standards Standard-frequency Transmissions

A continuous 24-hour service from the Bureau's radio stations WWV near Washington, D.C., and WWVH (Hawaii), comprises standard radio and audio frequencies, as listed in Table 14.21. At least

Table 14.21

Carrier frequency, ± 1 in 10^8 (Mc/s)	*Power* (kW)	*Modulation frequency,* ± 1 in 10^8 (c/s)
2·5 5	0·7 8	440
10 15 20 25	9 9 8·5 or 0·1 0·1	440 and 4,000

Minutes past each hour	*Type of modulation*
0–4, 5–9, 10–14, 15–19, 20–24, 25–29, 30–34, 35–39, 40–44, 45–49, 50–54, 55–59	Standard a.f. as above, plus 5-millisecond 1,000-c/s pulses at intervals of 1 second $\pm$ 1 μsec, audible as faint ticks. The 59th pulse in every minute is omitted.
4–5, 9–10, 14–15, 24–25, 34–35, 39–40, 44–45, 54–55	A.f. interrupted for 1 minute $\pm$ 1 in 10^8. WWV in morse.
29–30, 59–60	Station announcement by voice
19–20, 49–50	Radio propagation signals, as below

one, and often several, of the transmissions can usually be received anywhere in the world at any time.

The propagation signals, in morse, consist of series of the letter:

(1) " W ", meaning that radio propagation disturbance exists, especially on transmission paths crossing the North Atlantic; or

(2) " N ", meaning no warning; or

(3) " U ", meaning conditions unstable.

Propagation disturbance is characterized by flutter and rapid fading on the normal frequencies used at that time, or by complete blackout of signals.

Digits following the propagation letters indicate conditions expected during the next 12 hours, according to Table 14.22. The warnings do not apply to sudden ionospheric disturbances, which are unpredictable.

Propagation effects, such as Doppler effect, may at times cause slight fluctuations in the frequencies as received; the average frequency received is, however, as accurate as that sent.

Time intervals of 1 minute or more, marked by stopping and starting of a.f. modulation, are accurate to 2 in 10^8. Announced times are in E.S.T., which is 5 hours behind G.M.T.

It is, of course, easy to separate such diverse audio frequencies as 440 and 4,000 c/s by means of a simple filter; but for the frequency-comparison methods of Sec. 10.10 it is not necessary to do so, because

Table 14.22

Digit	*Propagation condition*
1	impossible
2	very poor
3	poor
4	fair to poor
5	fair
6	fair to good
7	good
8	very good
9	excellent

only one of the patterns can be made stationary on the screen at a time.

Inquiries should be addressed to National Bureau of Standards, Boulder Laboratories, Boulder, Col., U.S.A.

14.38. TEST DISK AND TAPE RECORDS

Disks

Table 14.23 supplies data on test records. Although 78 r.p.m. record players are obsolescent, test records for this speed are retained in the list because besides the primary purposes of testing record-reproducing systems they can also be used—with an appropriate pick-up and turntable—as sources of signals for other purposes. For example, JH.138 is a substitute for two audio signal generators and mixing network for intermodulation tests.

Where the frequency characteristic is given as " B.S." it conforms to the C.C.I.R. standards set out in B.S. 1928 : 1960 (see ref. below) and shown in Fig. 14.47, except that all recordings above 10 kc/s are halved in amplitude (i.e., – 6 dB) to avoid excessive amplitudes of cut, and response figures must accordingly be doubled over this frequency range. The curves in Fig. 14.47 are plotted to the equation

$$N(\text{dB}) = 10 \log (1 + \omega^2 t_1^2) - 10 \log \left(1 + \frac{1}{\omega^2 t_2^2}\right) + 10 \log \left(1 + \frac{1}{\omega^2 t_3^2}\right)$$

where N is the number of dB with reference to the level at 1 kc/s (given

Table 14.23

Supplier	*Number and Material*	*Speed r.p.m.*	*Diameter in.*	*Groove*	*Frequency Charac.*	*Ref. level cm/sec (at 1 kc/s)*	*Side*	*Recording*
Cosmocord Ltd., Eleanor Cross Road, Waltham Cross, Herts.	—	45	7	fine mono.	B.S.	1·65 (+ 4·4 dB)	1	Constant tones: 10, 9, 8, 7, 6, 5, 4, 3, 2, 1 kc/s; 500, 250, 100, 50 c/s
							2	Extracts from five musical records
Decca Records Co. Ltd., Decca House, Albert Embankment, London, S.E.1., & Decca dealers	LXT.5346	33⅓	12	fine mono.	B.S.	1·2 (+ 1·6 dB)	1	Constant tones: 1, 18. 16, 14, 12, 10, 8, 7, 6, 5, 4, 3, 2, 1·5, 1 kc/s; 700, 500, 300, 200, 150, 100, 80, 60, 50, 40, 30 c/s
							2	Gliding tone: 18 kc/s to 30 c/s, with breaks every kc/s and at 500, 250, 120, 60 c/s
	SXL.2057	33⅓	12	fine stereo	B.S.	1·0	1	Left-hand channel; 12, 10, 8, 6, 4, 2, 1 kc/s; 500 250, 125, 60, 40 c/s
							2	Right-hand channel: ditto. Crosstalk better than − 20 dB
E.M.I. Records Ltd., 3 Abbey Road, London, N.W.8	JGS 81 (vinyl or shellac)	78	12	coarse mono.	B.S.	1·78 (+ 5 dB)	1 only	* Constant tones: 1, 18, 16, 14, 12, 10, 8, 6, 5, 4, 3, 2, 1 kc/s; 700, 400, 200, 110, 60, 30 c/s
	SR.21 (vinyl)	78	12	coarse mono.	See last column	3·85 (+ 11·7 dB)	1	Constant tones: 20, 18, 16, 14, 12, 10, 8, 6, 4, 2 kc/s (+ 0·3 dB); 1 kc/s (0 dB); 400 c/s (− 1·1 dB); 160 c/s (− 5·1 dB); 70 c/s (− 11·2 dB)
							2	Constant tones: 19, 17, 15, 13, 11, 9, 7, 5, 3, 1 kc/s (0 dB); 700 c/s (− 0·5 dB); 250 c/s (− 3·0 dB); 100 c/s (− 8·5 dB); 50 c/s (− 14 dB)

Supplier	*Number and Material*	*Speed r.p.m.*	*Diameter in.*	*Groove*	*Frequency Charac.*	*Ref. level cm/sec (at 1 kc/s)*	*Side*	*Recording*
E.M.I. Records Ltd., 3 Abbey Road, London, N.W.8	JH.138 (vinyl)	78	10	coarse mono.	See last column	1·0	1	Eleven bands of 400 c/s and 4 kc/s together; first band at + 22·5 dB and + 10·5 dB respectively; each subsequent band 2 dB below previous one
							2	Ditto, but 60 c/s and 2 kc/s, beginning at + 8·6 dB and + 10·3 dB respectively
E.M.I. Records Ltd., 20, Manchester Square, London, W.1., & E.M.I. dealers	TCS.101 (vinyl)	33⅓	12	fine stereo	B.S.	1·0	1	* First and last bands, lateral cut at 1 kc/s. Others, 20, 18, 16, 14, 12, 10, 8, 6, 5, 4, 3, 2, 1 kc/s; 700, 400, 200, 100, 60, 30 c/s, each successively left and right hand channel 45° cut
							2	Ditto
	TCS.102 (vinyl)	33⅓	12	fine stereo	B.S.	1·0	1	* Gliding tone, left hand channel; 20 kc/s to 30 c/s, preceded by 1 kc/s
							2	Ditto, right hand channel
	TCS.104	33⅓	12	fine mono.	B.S.	1·0	1	* Constant tones as TCS.101, but all lateral cut
							2	Gliding tone, 20 kc/s to 30 c/s, lateral cut, preceded by 1 kc/s
	TCS.105	33⅓	12	vertical	B.S.	1·0	1 2	* Constant tones as TCS.101, but all vertical cut Gliding tone, 20 kc/s to 30 c/s, vertical cut

* Recorded announcements of frequencies, etc.
All dB figures are relative to 1 cm/sec

Table 14.24

Supplier	Number	Material	Speed in/sec	Spool	Track	Recorded level	Frequencies for azimuth adjt., kc/s	Recording	Price £ s d
Ampex (G.B.) Ltd., Arkwright Road, Reading, Berkshire	4494	Mylar	15	NAB/BSI, 5 in.	Full	—	15	14 frequencies from 50 c/s to 15 kc/s	6 15 0
	5563	Mylar	7½	NAB/BSI, 5 in.	Full	—	10	12 frequencies from 50 c/s to 10 kc/s	6 15 0
	6000	Mylar	3¾	NAB/BSI, 5 in.	Full	—	7·5	13 frequencies from 50 c/s to 7·5 kc/s	6 15 0
	6874	Mylar	3¾	NAB/BSI, 5 in.	Full	—	7·5	10 frequencies from 50 c/s to 7·5 kc/s; blank length for bias adjustment	6 15 0
Basf Chemicals Ltd., 5a Gillespie Road, London, N.5	DIN 38	PVC	15	BSI, 7 in.	Full	20 dB below peak	10	14 frequencies from 30 c/s to 15 kc/s; blank length for bias adjustment	8 0 0
	DIN 19	PVC	7½	BSI, 5¾ in.	Full	20 dB below peak	10	13 frequencies from 30 c/s to 12 kc/s; blank length for bias adjustment	8 0 0
	DIN 9	PVC	3¾	BSI, 5in.	Full	20 dB below peak	8	11 frequencies from 30 c/s to 8 kc/s; blank length for bias adjustment	8 0 0
E.M.I. Electronics Ltd. (Inst. & Maint. Div.), Hayes, Middlesex	SRT 16	Acetate	7½ (25 frames per sec)	BSI, 7 in.	Centre and edge (B.S. 291)	—	10	14 frequencies from 40 c/s to 10 kc/s; 3 kc/s for wow and flutter check; percussion recording for transients; 1 kc/s at maximum level; erased length for noise check	25 0 0
E.M.I. Sales & Service Ltd. (Rec. Equip. Div.), Hayes, Middlesex	SRT 11	PVC	30	Cont./BSI, 7 in.	Full	12 dB below peak	15	15 frequencies from 40 c/s to 15 kc/s; 3 kc/s for wow and flutter check; 50 c/s stroboscope for speed check	10 0 0
	SRT 12	PVC	15	Cont./BSI, 7 in.	Full	12 dB below peak	15	15 frequencies from 40 c/s to 15 kc/s; 3 kc/s for wow and flutter check; 50 c/s stroboscope for speed check	10 0 0

Supplier	Number	Material	Speed, in/sec	Spool	Track	Recorded level	Frequencies for azimuth adjt., kc/s	Recording	Price £ s d
E.M.I. Sales & Service Ltd. (Rec. Equip. Div.), Hayes, Middlesex	SRT 13	PVC	7½	Cont./BSI, 7 in.	Full	12 dB below peak	12	12 frequencies from 40 c/s to 10 kc/s; 3 kc/s for wow and flutter check; 50 c/s stroboscope for speed check	10 0 0
	SRT 14	PVC	3¾	BSI, 7 in.	Full	12 dB below peak	12	12 frequencies from 40 c/s to 10 kc/s; 3 kc/s for wow and flutter check; 50 c/s stroboscope for speed check	10 0 0
E.M.I. Ltd. (Tape Record Dept.), Hayes, Middlesex	TBT 1	PVC	7½	BSI, 3 in.	Full	Arbitrary	8	8 kc/s for 60 sec for azimuth adjustment; 1 kc/s for 30 sec for level adjustment; 40, 60, 110, 200, 500 c/s and 1, 2, 4, 6, 8, 10 kc/s, each for 15 sec	3 0 0
Minn. Mining & Manf. Co. Ltd., 3M House, Wigmore Street, London, W.1	—	PVC	7½	BSI, 3¼ in.	Half	12 dB below peak (10 gauss at 1 kc/s)	7·5	11 frequencies from 40 c/s to 10 kc/s on track 1; alignment check on track 2	1 19 6
Tutchings Electronics Ltd., 14 Rook Hill Road, Friars Cliff, Christchurch, Hampshire	1	PVC	7½	BSI, 3¼ in.	Half	12 dB below peak (10 gauss at 1 kc/s)	7·5	11 frequencies from 40 c/s to 10 kc/s on track 1; alignment on track 2	1 19 6
	2	PVC	3¾	BSI, 3¼ in.	Half	− 12 dB (15 gauss)	5	9 frequencies from 40 c/s to 10 kc/s on track 1; alignment on track 2	1 19 6
	3	PVC	7½	BSI, 3¼ in.	Half	12 dB below peak	—	White noise, divided into 25 ⅓-octave bands	1 19 6
	4	PVC	7½	BSI, 3¼ in.	Half	12 dB below peak	—	White noise, divided into 7 1-octave bands on track 1; full white noise on track 2	1 19 6
	5	PVC	—	BSI, 3¼ in.	Half	12 dB below peak	—	Full-range white noise for azimuth adjustment	1 19 6

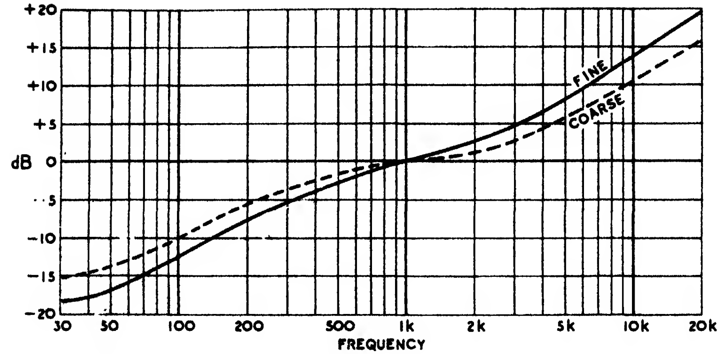


Fig. 14.47—British Standard recording frequency characteristics for coarse-groove (dotted line) and fine-groove (full line) disks

in the next column of the Table), ω is $2\pi f$, and t_1, t_2 and t_3 are the time constants of the recording tone control circuits: 50, 450 and 3,180 for coarse-groove records and 75, 318 and 3,180 for fine-groove records, in all cases multiplied by 10^{-6}. B.S. 1928 allows $\pm$ 2 dB tolerance, but most of the test records are much better than this.

The colour code for coarse-groove equipment is green, and for fine-groove red. Coarse grooves are used only for 78 r.p.m. disks and for 16-in. 33⅓ r.p.m. transcription recordings.

Refs: B.S. 1928 : 1960, *Gramophone Records, Transcription Disk Recordings and Disk Reproducing Eqiupment* (B.S.I.).
" Recording and Reproducing Equalisation ", by P. J. Guy. *E.E.*, Jan. 1961.

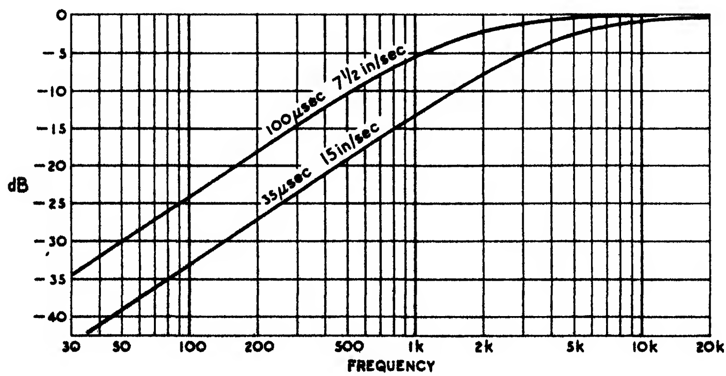


Fig. 14.48—C.C.I.R. Standard recording frequency characteristics for magnetic tapes

Tapes

Table 14.24 is based on information compiled by D. Aldous and G. Balmain. Frequencies, etc., are announced on the tapes. The SRT tapes are professional laboratory recordings conforming to C.C.I.R. standards set out in B.S. 1568 : 1953, Amendment 1. The standard frequency characteristics for recording are shown in Fig. 14.48. Full details of the tape specifications, and of the precautions to be observed in handling, storing and using these tapes, are obtainable from the makers.

The TBT.1, obtainable from the Tape Record Department, is an inexpensive tape intended mainly for amateur use. It is recorded full-track, so is suitable for checking either mono or stereo reproducers. The frequency characteristic is to C.C.I.R. standard.

Ref: B.S. 1568 : 1960, *Magnetic Tape Sound-Recording and Reproduction for Programme Interchange* (B.S.I.).

14.39. COLOUR CODES

Fixed Resistors and Capacitors

The following are specified by the British Radio Industry Council. American coding is the same as regards values, but differs in some of the other details.

Table 14.25

Colour	*A, B, C: Value (Ω or pF)*				*D: Tolerance*			*E: Temp. coefficient per* 10^6 *per °C*
	A: 1st figure	*B: 2nd figure*	*C: Multiplier*		*Resistors*	*Ceramic capacitors*		
			Resistors	*Capacitors*		*10 pF or less*	*> 10 pF*	
black	—	0	1	1	—	2 pF	±20%	0
brown	1	1	10	10	± 1%	0·1 pF	± 1%	−30
red	2	2	10^2	10^2	± 2%	—	± 2%	−80
orange	3	3	10^3	10^3	—	—	±2·5%	−150
yellow	4	4	10^4	10^4	—	—	—	−220
green	5	5	10^5	—	—	0·5 pF	± 5%	−330
blue	6	6	10^6	—	—	—	—	−470
violet	7	7	10^7	—	—	—	—	−750
grey	8	8	10^8	10^{-2}	—	0·25 pF	—	+30
white	9	9	10^9	10^{-1}	—	1 pF	±10%	+100
silver	—	—	10^{-2}	—	±10%	—	—	—
gold	—	—	10^{-}	—	± 5%	—	—	—

Standard ± tolerances for resistors are:

Wire-wound: **1%, 2%, 5%, 10%.**

Composition, grade **1: 1%, 2%, 5%.**
grade **2: 5%, 10%, 20% (20% is indicated by no D colour).**

Grade 1 ("high-stability") composition resistors are distinguished by a salmon-pink fifth ring or body colour.

Ref: B.S. 1852: 1952. ***Colour Code for Fixed Resistors for Telecommunication Purposes*** **(B.S.I.).**

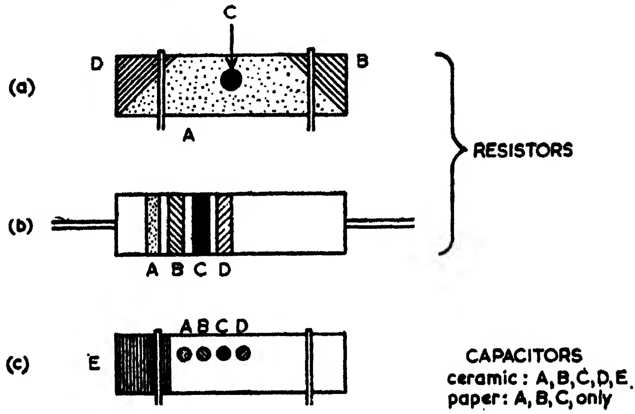


Fig. 14.49—Standard colour markings of resistors and capacitors; for key, see Table 14.25. An alternative to the spot C in (a) *is a band. A salmon-pink fifth band in* (b) *indicates Grade 1*

14.40. WIRE TABLES

The accompanying wire tables, which cover almost every practical need, are published by kind permission of the compiler, C. R. Cosens, M.A., and of the Editor of *The Journal of Scientific Instruments*, in the April 1937 issue of which they first appeared. They have been extended by the present author to include the last three gauges, which, however, because of their extreme fragility, are not highly recommended.

The following are some extracts from Mr. Cosens's explanatory remarks.

" The results given are fair averages, and different samples should not vary therefrom by more than a few per cent. In practical use, any apparent improvement on 'slide-rule accuracy' is illusory; for example, a change of temperature of 2·5° C alters the resistance of a copper wire by about 1%, and although resistance materials have much smaller temperature coefficients of resistance, they vary from sample to sample with unavoidable small changes of composition and of wire diameter, due to wear of the drawing dies. If great accuracy is required, it is advisable to determine the relation between length of wire and resistance by experiment, and then to mark the result on the reel.

" It should be remembered that when manganin is annealed after winding there will be an appreciable drop of resistance, so that a coil should be wound about 1 or 2% high and adjusted after annealing. Annealing is not necessary for constantan; though desirable for very accurate work, the change in resistance is very small.

" The columns for D.S.C. wire can also be used for finding the length on a reel of enamel and single silk-covered wire, and the error will be small and on the safe side.

"The tables give the number of turns that can be wound in 1 sq. cm. of a multi-layer coil such as a transformer coil. This has been calculated with no allowance for 'bedding', i.e., on the assumption that the wires will lie as shown in Fig. 14.50, not as in Fig. 14.51. Although Fig. 14.51 is possibly nearer the truth when a wire winding machine is used, experience shows that when wire is wound by hand in a lathe,

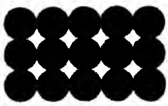


Fig. 14.50

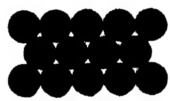


Fig. 14.51

without winding absolutely even layers for the finer sizes (which is incredibly tedious), but with reasonable care to keep the winding as level as possible, the values given in the table do fairly represent the number of turns that can be got into the available space. Additional space must of course be allowed for the insulating bobbin, or for the tape if former-wound.

"This information is given for five different wire coverings. A single layer of insulation, whether silk, cotton, or enamel, is not recommended for hand winding; enamel alone is very satisfactory when used on a winding machine, but when wound by hand there is risk of scratching the enamel or pinching the wire and causing a short-circuited turn. Enamel and single silk (E. & S.S.) or enamel and single cotton (E. & S.C.) are very satisfactory; E. & S.S. can be specially recommended for small transformers, it takes up no more space than double silk (sometimes less), and it is cheaper. For thicker wires, E. & S.C. takes up much less space than double cotton, and is no more expensive. Both these coverings are excellent from the point of view of insulation: the cotton or silk makes a soft bed for the turns of wire so that the enamel is not easily injured, while the enamel gives high insulation resistance even if a 'dry' coil without varnish is used in a damp place.

"Where the errors due to thermal e.m.fs cannot be neglected, it is necessary to use manganin (copper 84%, manganese 12%, nickel 4%), but some little experience is needed to work this material, so that, apart from the expense, one does not use it if it can be avoided. Manganin cannot be soft-soldered, but must be silver-soldered at a red heat with a blowpipe, a process requiring some dexterity with a fine wire. It is also necessary to have some form of annealing oven, preferably electric, which must have thermostatic control to keep the temperature correct to within a degree or two, as it is necessary after winding to coat the wire with shellac varnish to protect it from oxidization, and then to anneal for at least 24 hours at a temperature not below

Table 14.26

	Copper			Constantan (or Eureka)			Manganin			
S.W.G.	*Ohms per km*	**Metres per 1 oz reel* D.S.C.	**Metres per 1 lb reel* D.C.C.	*Cm per ohm*	*Ohms per 1 oz reel* D.S.C.	*Ohms per 1 lb reel* D.C.C.	*Cm per ohm*	*Ohms per 1 oz reel* D.S.C.	*Ohms per 1 lb reel* D.C.C.	S.W.G.
12	3·09	0·57	9·1	1,140	0·050	0·80	1,300	0·044	0·70	12
13	3·95	0·73	11·5	890	0·082	1·32	1,020	0·072	1·15	13
14	5·22	0·97	15·2	680	0·144	2·3	770	0·126	2·0	14
15	6·45	1·20	18·7	550	0·22	3·5	630	0·192	3·1	15
16	8·16	1·52	24	430	0·35	5·6	490	0·31	4·9	16
17	10·66	1·98	30	330	0·60	9·6	380	0·52	8·4	17
18	14·51	2·7	41	240	1·10	17·7	280	0·97	15·4	18
19	20·9	3·8	58	169	2·3	36	193	1·99	32	19
20	25·8	4·7	71	137	3·5	55	156	3·0	48	20
21	32·6	6·0	90	108	5·5	88	124	4·8	77	21
22	42·6	7·8	115	83	9·4	150	95	8·2	131	22
23	58·0	10·5	156	61	17·3	280	70	15·1	240	23
24	69·1	12·6	183	51	25	400	58	22	350	24
25	83·6	15·3	220	42	36	580	48	32	510	25
26	103·2	18·7	260	34	55	880	39	48	770	26
27	124·3	22	310	28	79	1,270	33	69	1,110	27
28	152·6	28	380	23	119	1,900	26	104	1,670	28
29	180·8	32	430	19·5	167	2,700	22	146	2,300	29
30	217	39	500	16·2	240	3,800	18·6	210	3,400	30

* These columns may also be used for Constantan and Manganin.

S.W.G.	Copper			Constantan (or Eureka)			Manganin			S.W.G.
	Ohms per km	**Metres per 1 oz reel* D.S.C.	**Metres per 1 lb reel* D.C.C.	*Cm per ohm*	*Ohms per 1 oz reel* D.S.C.	*Ohms per 1 lb reel* D.C.C.	*Cm per ohm*	*Ohms per 1 oz reel* D.S.C.	*Ohms per 1 lb reel* D.C.C.	
31	249	44	560	14·2	310	5,000	16·2	270	4,440	31
32	287	51	640	12·3	410	6,660	14·1	360	5,800	32
33	334	59	730	10·6	560	9,000	12·1	490	7,900	33
34	395	69	830	8·9	780	12,400	10·2	680	10,900	34
35	474	83	1,010	7·5	1,100	17,800	8·5	970	15,600	35
36	579	100	1,210	6·1	1,640	26,200	7·0	1,430	23,000	36
37	723	123	1,400	4·9	2,500	40,000	5·6	2,220	35,000	37
38	929	158	1,740	3·8	4,100	66,000	4·3	3,600	58,000	38
39	1,236	210	2,200	2·9	7,200	115,000	3·3	6,300	101,000	39
40	1,451	240	2,400	2·4	9,800	157,000	2·8	8,600	137,000	40
41	1,727	280	—	2·0	13,600	—	2·3	11,900	—	41
42	2,090	330	—	1·69	19,600	—	1·93	17,100	—	42
43	2,580	390	—	1·37	29,000	—	1·56	25,000	—	43
44	3,270	480	—	1·08	44,000	—	1·24	39,000	—	44
45	4,260	570	—	0·83	69,000	—	0·95	60,000	—	45
46	5,800	730	—	0·61	120,000	—	0·70	105,000	—	46
47	8,360	1,020	—	0·42	220,000	—	0·48	196,000	—	47
48	13,050	1,480	—	0·27	550,000	—	0·31	480,000	—	48
49	23,200	2,380	—	0·15	1,550,000	—	0·17	1,350,000	—	49
50	33,400	3,200	—	0·105	3,000,000	—	0·12	2,650,000	—	50

* These columns may also be used for Constantan and Manganin.

Table 14.27

S.W.G.	*Turns per cm length of single-layer coil*					*Turns per sq. cm. section of solid coil*					Dia-meter (in) (bare)	Cross-section (sq. mm)	Dia-meter (mm) (bare)	S.W.G.
	D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	*Enamel*	D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	*Enamel*				
12	—	3·3		3·4	3·6	—	11·1	—	11·7	13·3	0·104	5·48	2·64	12
13	—	3·7		3·8	4·1	—	13·8	—	14·6	16·8	0·092	4·29	2·34	13
14	—	4·2	—	4·3	4·7	—	17·5	—	18·7	22	0·080	3·24	2·03	14
15	—	4·7	—	4·7	5·2	—	22	—	23	27	0·072	2·63	1·83	15
16	5·8	5·2	5·6	5·3	5·8	33	27	32	28	34	0·064	2·07	1·63	16
17	6·6	5·8	6·4	6·0	6·7	43	33	41	36	45	0·056	1·59	1·42	17
18	7·7	6·7	7·5	6·8	7·8	60	44	56	46	60	0·048	1·167	1·22	18
19	9·2	7·7	8·8	7·9	9·2	84	60	78	62	85	0·040	0·811	1·02	19
20	10·1	8·4	9·7	8·6	10·2	102	70	94	73	103	0·036	0·657	0·92	20
21	11·2	9·2	10·8	9·4	11·4	127	84	116	88	130	0·032	0·519	0·81	21
22	12·7	10·1	12·1	10·4	12·9	161	102	147	107	167	0·028	0·397	0·71	22
23	14·6	11·6	13·9	12·5	15·0	210	134	193	156	220	0·024	0·292	0·61	23
24	15·7	12·3	15·3	13·3	16·2	250	151	230	178	260	0·022	0·245	0·56	24
25	17·1	13·1	16·9	14·6	18·1	290	172	280	210	330	0·020	0·203	0·51	25
26	19·2	14·1	18·7	15·7	19·9	370	198	350	250	400	0·018	0·1642	0·46	26
27	21	14·9	20	17·2	22	430	220	420	300	480	0·0164	0·1363	0·42	27
28	23	15·9	22	18·5	24	520	250	500	340	580	0·0148	0·1110	0·38	28
29	24	16·7	24	19·6	26	600	280	580	380	670	0·0136	0·0937	0·35	29
30	26	17·6	26	21	29	700	310	700	450	840	0·0124	0·0779	0·32	30

S.W.G.	Turns per cm length of single-layer coil					Turns per sq. cm. section of solid coil					Dia-meter (in) (bare)	Cross-section (sq. mm)	Dia-meter (mm) (bare)	S.W.G.
	D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel	D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel				
31	28	18·2	28	22	31	780	330	780	490	940	0·0116	0·0682	0·29	31
32	30	18·9	30	23	33	880	360	880	540	1,080	0·0108	0·0591	0·27	32
33	31	19·7	32	24	35	990	390	990	590	1,240	0·0100	0·0507	0·25	33
34	34	21	35	26	39	1,170	420	1,190	670	1,490	0·0092	0·0429	0·23	34
35	36	23	37	29	42	1,300	510	1,380	860	1,750	0·0084	0·0358	0·21	35
36	39	24	40	31	46	1,520	560	1,610	980	2,100	0·0076	0·0293	0·19	36
37	42	25	44	—	51	1,790	620	1,910	—	2,500	0·0068	0·0234	0·17	37
38	46	26	49	—	56	2,100	690	2,400	—	3,200	0·0060	0·01824	0·15	38
39	51	28	55	—	66	2,600	770	3,000	—	4,300	0·0052	0·01370	0·13	39
40	54	29	58	—	72	2,900	810	3,400	—	5,100	0·0048	0·01167	0·12	40
41	60	—	62	—	79	3,600	—	3,800	—	6,200	0·0044	0·00981	0·11	41
42	64	—	66	—	86	4,000	—	4,300	—	7,300	0·0040	0·00811	0.10	42
43	68	—	73	—	96	4,600	—	5,300	—	9,200	0·0036	0·00657	0·09	43
44	73	—	79	—	106	5,300	—	6,200	—	11,300	0·0032	0·00519	0·08	44
45	79	—	86	—	119	6,200	—	7,300	—	14,200	0·0028	0·00397	0·07	45
46	86	—	96	—	141	7,300	—	9,200	—	19,800	0·0024	0·00292	0·06	46
47	94	—	109	—	171	8,800	—	12,000	—	29,000	0·0020	0·00203	0·05	47
48	108	—	—	—	206	10,800	—	—	—	43,000	0·0016	0·00130	0·04	48
49	—	—	—	—	280	—	—	—	—	79,000	0·0012	0·00073	0·03	49
50	—	—	—	—	328	—	—	—	—	107,000	0·0010	0·000507	0·025	50

130° C. If the temperature does not exceed 130° C, the manganin does not anneal; if it exceeds 140° C, the silk insulation will be charred. If the shellac coat does not completely cover the wire, oxidization takes place and the resulting coil will have a large temperature coefficient (when annealed properly without access of oxygen the temperature coefficient is very small, sometimes even zero or negative).

" The cross-section in square millimetres is convenient for determining the current capacity. Drysdale and Jolley give the current density employed in instrument work as between 1 and 4 A per sq. mm. For small mains transformers of 100 W or so, higher current densities could be employed without dangerous overheating, but the determining factor is usually the voltage drop allowable on full load, and this results in demanding a current density of from 1 to 2 A per sq. mm. only."

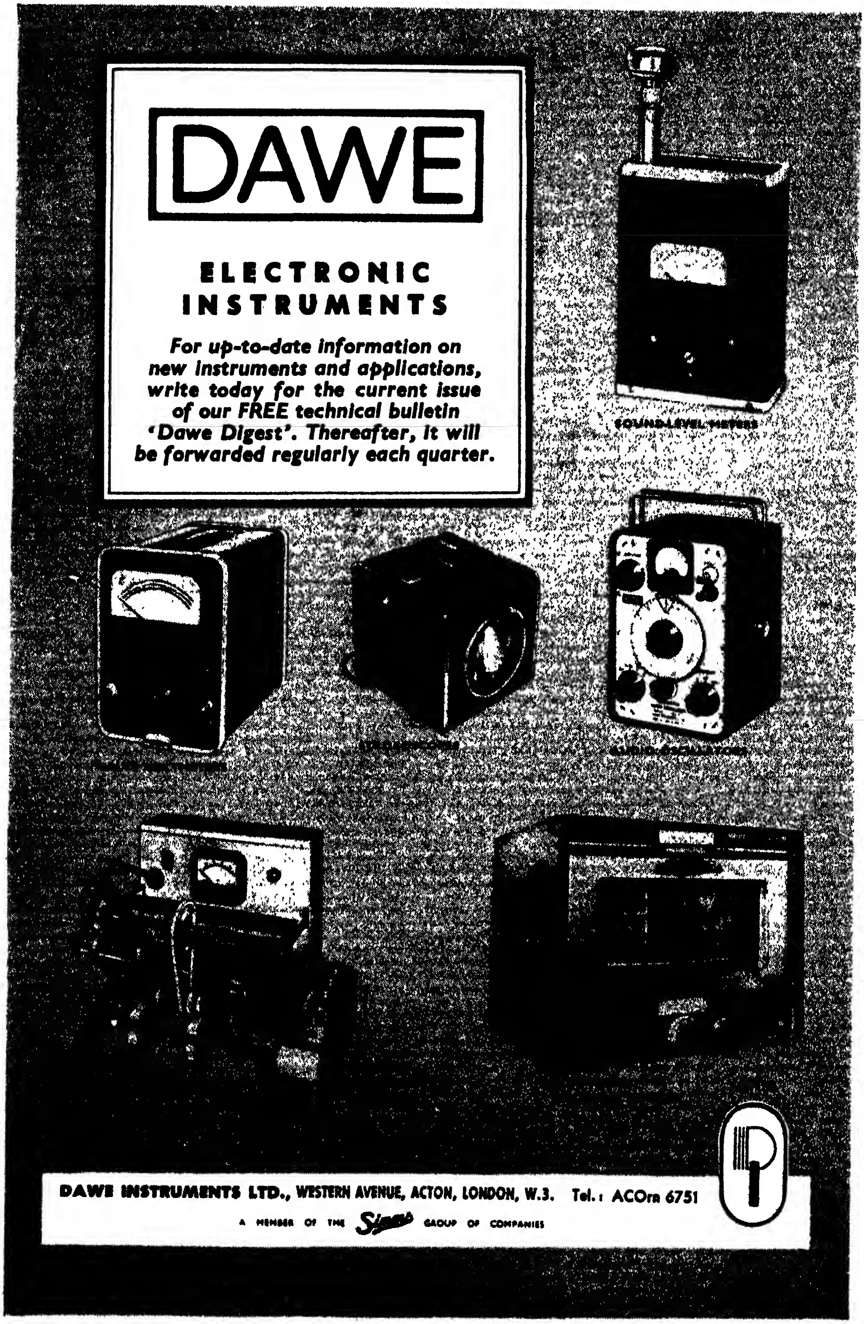
Karma alloy has the very low temperature coefficient of Eureka, with 2·8 times its resistivity, and is available in very small gauges, down to 0·0006 in. dia. (equivalent to 58 s.w.g.) giving 6·6 kΩ per yard. But it is even more difficult to solder than manganin and requires a special flux. One such can be made up as follows:

Aniline (commercial)	51 c.c.
Phosphoric Acid (S.G. 1·75)	34 c.c.

The acid is added slowly to the aniline, care being taken to keep the container cool. The aniline phosphate separates as a solid, which is then mixed to a paste with ethylene glycol or butyl cellosolve. There are one or two commercial fluxes of a similar character.

Index

The following are printed in italics:
(1) Names of persons whose work is referred to.
(2) Names of manufacturers of equipment.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