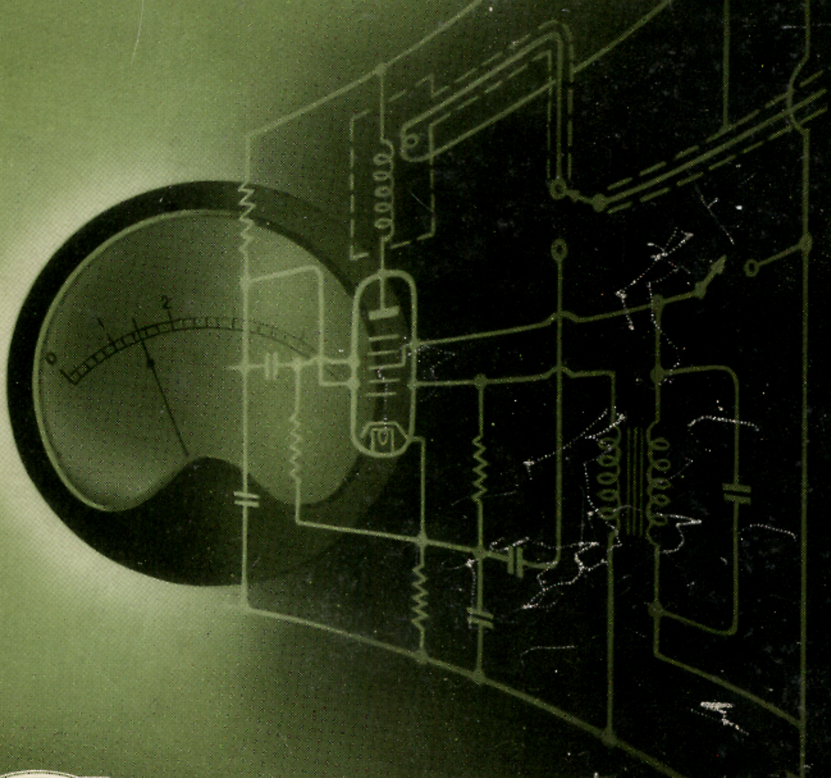
RADIO LABORATORY HANDBOOK



ASSOCIATED ILIFFE TECHNICAL BOOKS

M. G. SCROGGIE, B.Sc., M.I.E.E.

A "Wireless World" Publication

RADIO LABORATORY HANDBOOK

SIXTH EDITION

ILIFFE

RADIO LABORATORY HANDBOOK

THIS standard reference book on laboratory electronic techniques, written by a well-known consulting engineer, is intended for both professional engineers and amateurs. It first describes the layout and furnishing of an up-to-date laboratory, and then the various types of apparatus available. Both commercial instruments and improvised equipment are covered. Later chapters deal in detail with methods of making measurements and tests of every kind. Finally, a large section is devoted to general principles and reference material of everyday use to the radio engineer.

The book is particularly suitable for workers in sciences other than radio, such as physiology, in which electronic techniques are being increasingly used, because comparatively little radio experience is assumed and adequate explanations and references are given, while particular attention is given to the general principles of experimentation and interpreting results.

The large amount of information contained in the volume has been presented as concisely as possible and made easily accessible by the very comprehensive system of cross-references, table of contents and index. A special feature is the many carefully selected references to further information.

"Radio Laboratory Handbook" was originally published in 1938, and quickly gained recognition as a lucid and not-too-solemn practical manual filling the gap between "popular" home experimenters' literature and the advanced professional textbooks. In the intervening period, there have been extensive developments in techniques and equipment, with which subsequent editions have kept pace. This sixth edition has been almost entirely rewritten and greatly enlarged and is now presented in a new format. The usefulness of the text is enhanced by some 300 photographs, drawings and circuit diagrams.

BY THE SAME AUTHOR

FOUNDATIONS OF WIRELESS
(Fifth Edition)

RADIO LABORATORY HANDBOOK

By

M. G. SCROGGIE
B.Sc., M.I.E.E.
Consulting Radio Engineer

With 299 illustrations

SIXTH EDITION

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Preface

"You, lifting up this book to find out whether it is for dignified engineers, like yourself, and not just for amateurs (or, alternatively, for enthusiastic home experimenters, like yourself, and not just for dull professionals) are to be informed without delay that it is humbly dedicated to the use of both. Though the most gifted writer may fail to interest all of the people all the time, it is the present author's hope that he may succeed in interesting some of the people (that is, all those who experiment in radio, on however small or large a scale) some of the time."

That hope was expressed in the first edition. The various editions that have had to be produced since may perhaps be taken to indicate fulfilment. Evidently there has been some use for a book occupying the gap between the home experimenters' literature and the very solemn professional textbooks. During these fifteen years many new techniques have been developed. And whereas in 1938 there was so little choice of ready-made equipment that even where money was not the limiting factor it was necessary to construct much of what was needed, now the variety offered by the instrument trade is embarrassingly large. Meanwhile even the amateur has become more sophisticated, and his laboratory may be equipped in a manner that would once have been the envy of the professional.

All this being so, it was clear that mere revision would no longer do. Experimenting is a good deal less light-hearted than it used to be; rough approximations are less acceptable, and chapter and verse have to be quoted for everything. But although this sixth edition has been almost completely rewritten and considerably enlarged, it is hoped that it has not become so grave as to disappoint those readers who have been good enough to say they liked the original style. A most determined effort has been made to keep it within reasonable bounds by sorting out from the vast existing amount of instruments and methods those likely to be most generally useful—especially where funds are not too plentiful—and presenting the information concisely. The most important omission is the subject of microwaves, which now has a very large literature of its own. Account has, however, been taken of the importance of television.

To avoid having to repeat detailed descriptions of apparatus in every method employing it, the chapters on apparatus come first and then those on methods. General principles and reference material are stored at the end in Chapter 14. Any necessary link-up between these separate departments 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 any other units are used they are specified in square brackets; e.g.,

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

Symbols and abbreviations are British Standard, and are set out in detail in Secs. 14.1 and 2. The one Continental "intruder" is so useful that it is hoped that before long naturalization will be granted. It is the nanofarad (nF), equal to 1,000 pF or 0·001 μF. A very large proportion of the capacitances used in radio are of the order of 1 nF, and the 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 irritating and 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 itself to denote the main unit (F, H, or Ω), the values can be shown clearly on the most complicated circuit diagram.

Another usage, which has become more general since advocated in the first edition but 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 "d.c. voltage".

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 or article is vastly more important and useful than the volume number. Any seeming 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; the longer titles of journals are abbreviated as shown opposite. The 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 suggestions and proof-reading assistance given by Mr. C. R. Cosens of Cambridge University, as well as his encouragement to write the first edition, are gratefully acknowledged.

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

M.G.S.

Elstree Hill Laboratory,
Bromley, Kent.
February, 1954.

Full Titles of Journals

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. Insts.	Journal of Scientific Instruments.
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.

List of Publishers

Abbreviated Name	*Full Name and London Address*
B.S.I.	British Standards Institution, 2, Park Street, W.1.
Chambers	W. & R. Chambers, Ltd., 6, Dean Street, W.1.
Chapman & Hall	Chapman & Hall, Ltd., 37, Essex Street, W.C.2.
Cleaver-Hume	Cleaver-Hume Press, Ltd., 42a, South Audley Street, W.1.
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 & Sons, Ltd., Dorset House, Stamford Street, S.E.1.
Macdonald	Macdonald & Co. (Publishers), Ltd., 16, Maddox Street, W.1.
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.
Trader	Trader Publishing Co., Ltd., Dorset House, Stamford Street, S.E.1.

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 of radio 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 radio-manufacturing concerns, and on a smaller scale the same object is pursued by keen amateurs. Although making one's set because it is cheaper than buying it is no longer a valid motive, it is fascinating and instructive to compare different methods and to try one's own ideas.

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 investigate the properties of 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.

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 ". With this equipment the experimenter can study such things as the influence on v.h.f. propagation of time, season, weather, and astronomical phenomena such as sunspot activity. Although much of this work has been very thoroughly prosecuted by bodies such as the Radio Research Board, it has not been exhausted, particularly on the very short waves. These 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 full 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—oscillators, distortion meters, valve 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 signal generators that stopped short of the television frequencies find them very restricted to-day. Valve 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.

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 reception. In more than one case known to the author, large manufacturing firms eventually took over house property for laboratories.

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 usual huts 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. One reader recommends sheets of Windolite for this purpose, fixed so as to trap a layer of air between them and the window proper, and reinforced by Scotch tape to seal the draughts.

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

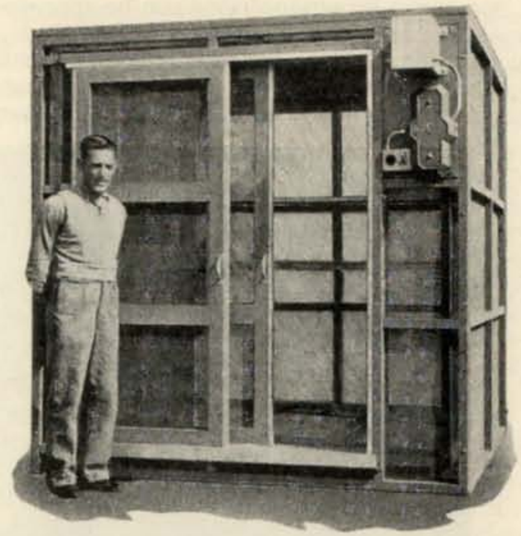


Fig. 2.1 — A completely screened test cabin. Note the mains interference filter and the 1 : 1 mains transformer for reducing risk of shock by providing an unearthed power supply.
(Belling & Lee, Ltd.)

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 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 on the ground keeps the feet thoroughly

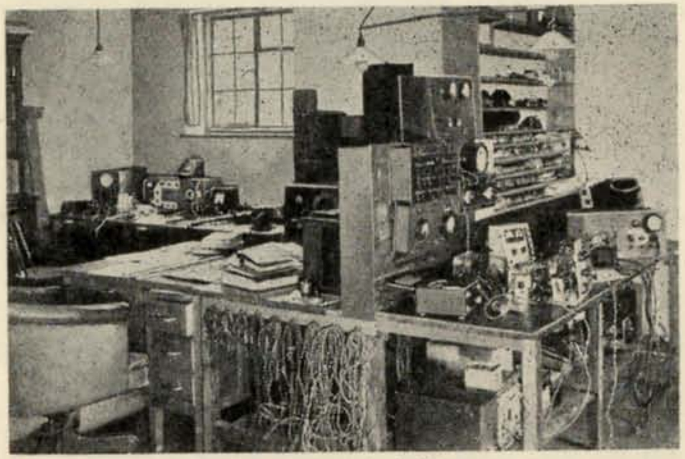


Fig. 2.2—In the author's laboratory

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 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

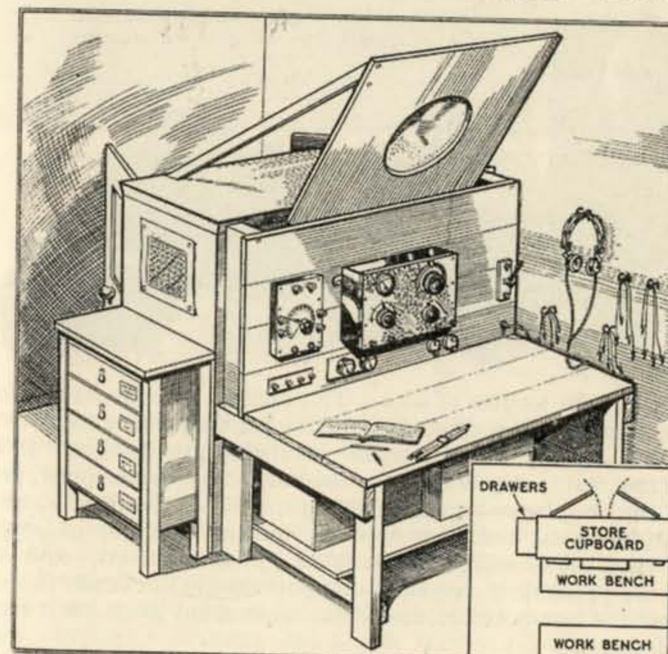


Fig. 2.3—Plan and perspective views of a compact laboratory layout, with a store cupboard and work-bench back to back. Note space between for concealed wiring, shelf below for batteries, etc., vertical mounting of apparatus, and hooks for connecting leads

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 near at hand, except the less-used gear which is kept on the other side. Fig. 2.3 shows a useful layout of this type. 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. 2.4) consists of a number of separate components—valve holders, transformers, etc.—fitted with terminals

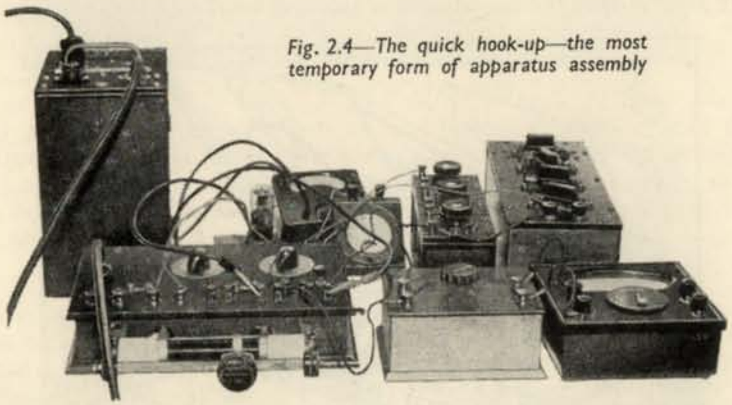


Fig. 2.4—The quick hook-up—the most temporary form of apparatus assembly

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

The Breadboard (Fig. 2.5) takes a little longer to set up and dismantle, and is not definitely a horizontal-space saver; but it is practically 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,

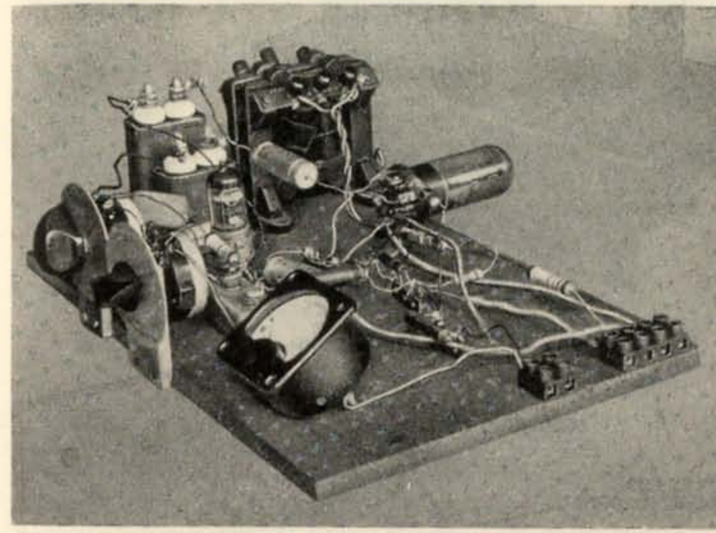


Fig. 2.5—An example of " breadboard " assembly

Fig. 2.6—Shrouded connector blocks are better than ordinary terminals for most purposes, temporary or permanent. This Belling-Lee type has the advantage of flexibility

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. 2.6, 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.

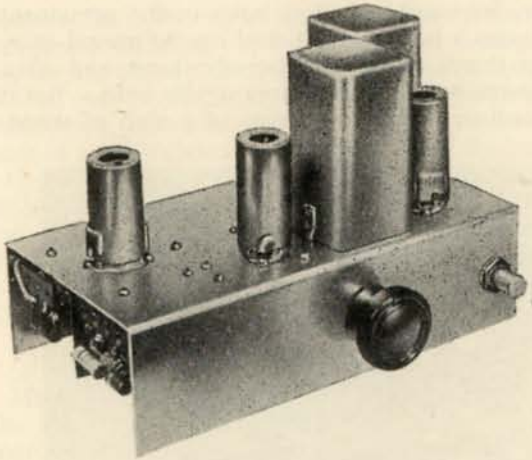


Fig. 2.7—An example of the chassis form of assembly

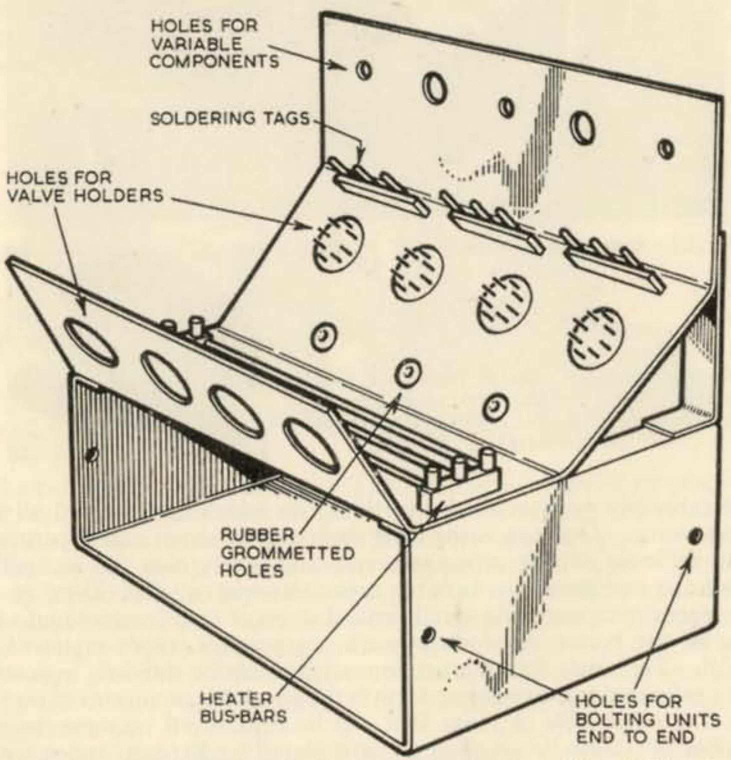
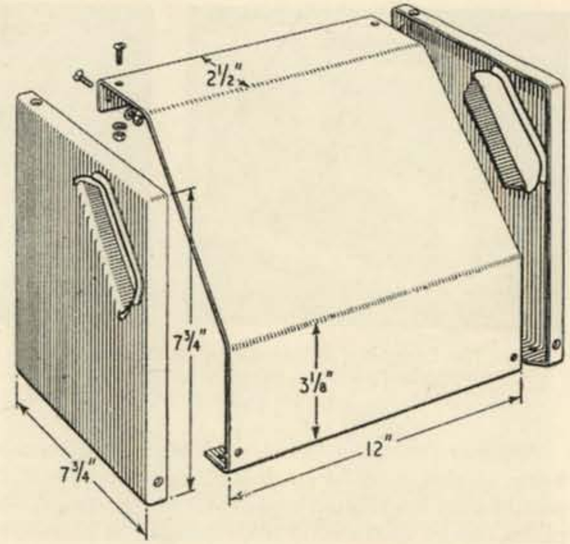


Fig. 2.8—The breadboard chassis—a labour-saving experimental layout of American design. The valves plug in from underneath

Fig. 2.9—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



If the board has $\frac{1}{2}$-in to $\frac{3}{4}$-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.*

The Chassis (Fig. 2.7), a familiar mount consisting of an inverted metal tray or shallow box, can be obtained from radio supply stores, or made up to size from metal sheet. If sloping sides are not objected to, it is also purchasable, tinned ready for soldering, from multiple stores, where it is better known as the baking tin.

Aluminium is very easy to work but difficult for making sound contacts; copper is good but expensive; steel is generally used for manufactured apparatus. The metal chassis is the most practical medium for screened construction; and, where a permanent assembly of a fairly adaptable nature is required, is preferable to the breadboard.

A hybrid type of assembly, the breadboard chassis,† can save an immense amount of time and trouble in all laboratories where many experimental valve circuits have to be tried. Fig. 2.8 is a sketch of the recommended form, but obviously it can be adapted to one's own requirements. Short wiring is combined with really good accessibility.

* " Fabricating Circuits on Plastic Breadboards ", by J. H. Bigbee. *Electronics*, September 1952, pp. 126–7.

† " Modern Breadboard Chassis ", by Snavely, Brown and Atanasoff. *Electronics*, July 1949, pp. 101–3.

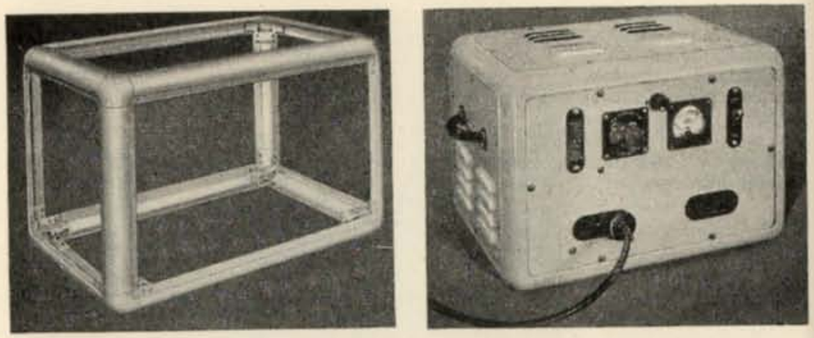


Fig. 2.10—Example of Widney-Dorlec construction, showing (left) framework made up from standard parts and (right) appearance with typical instrument in position. (Hallam Sleigh & Cheston, Ltd.)

Another form of chassis, useful for both temporary and permanent work, is that devised by B.B.C. engineers (Fig. 2.9); 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.

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 pieces, as shown in Fig. 2.10. Although easy to construct, the result is neat and professional.

* Hallam Sleigh & Cheston, Ltd., Bagot Street, Birmingham, 4.

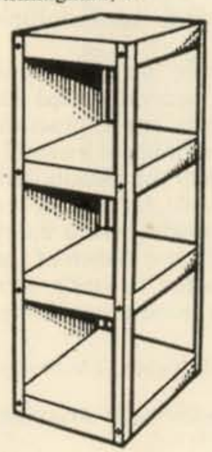


Fig. 2.11 (left)—Platform frame and rack for mounting several units in tiers

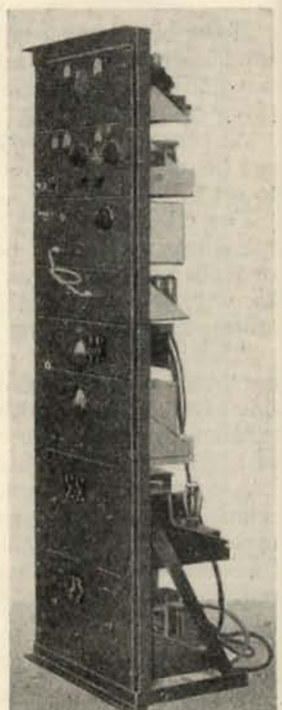
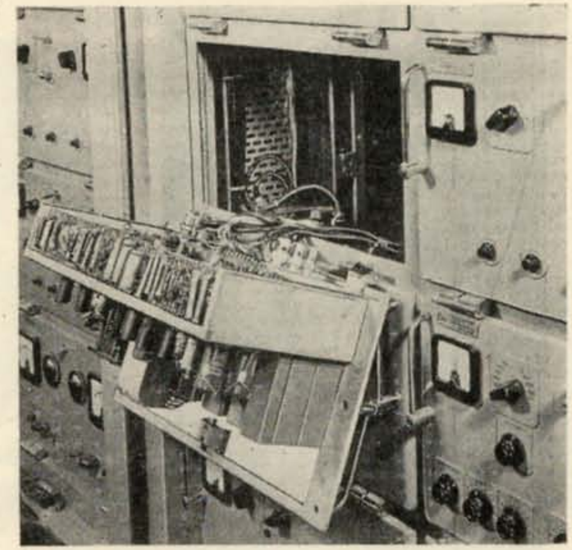


Fig. 2.12 (right)—Vertical rack mounting for panels comprising a large assembly

Fig. 2.13—Rack-mounted panel hinged for accessibility



The Platform Frame (Fig. 2.11) 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. 2.12). 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

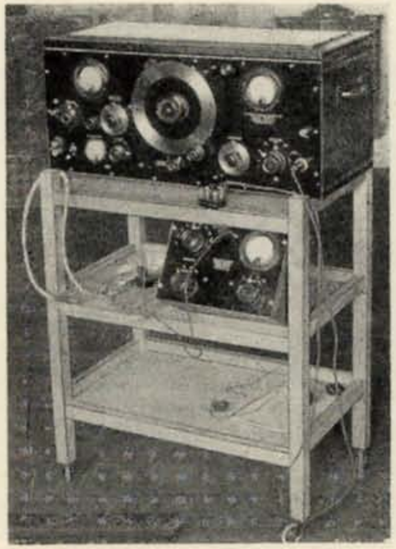


Fig. 2.14—Trolley mounting is particularly suitable for much-used bulky units

Fig. 2.15—Oscilloscope mounting trolley. (Nagard Ltd.)

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, as shown in Fig. 2.13. 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.

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. 2.14). 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. 2.15 shows a specialized form of mobile mounting for oscilloscope and power unit.

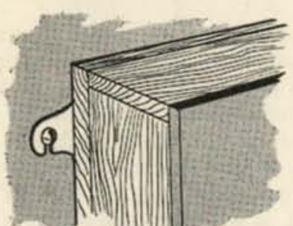


Fig. 2.16—Mirror plates screwed to box-form units enable them to be mounted on vertical surfaces

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

* Dexion, Ltd., 189, Regent Street, London, W.1.

hung one above another on the wall or backboard of the bench (Fig. 2.16).

For more information on general design of equipment, particularly for industrial laboratories, see an article by G. H. Hickling in *Electronic Engineering*, May 1952, pp. 236–8.

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 invariably shrink and warp, leaving crevices for small screws to hide in, so a sheet of thick linoleum or hardboard (such as Masonite) 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.

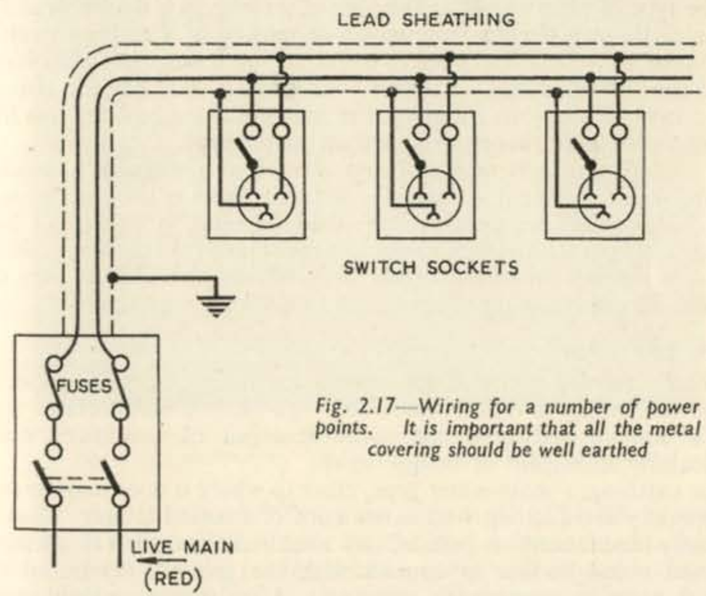


Fig. 2.17—Wiring for a number of power points. It is important that all the metal covering should be well earthed

Then there is wiring. An abundance of electric supply points should be provided, otherwise, when there is a soldering iron, a check receiver, a mains-driven oscillator, and a valve-heating transformer 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. 2.17). 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. 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.

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. Connections

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

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.

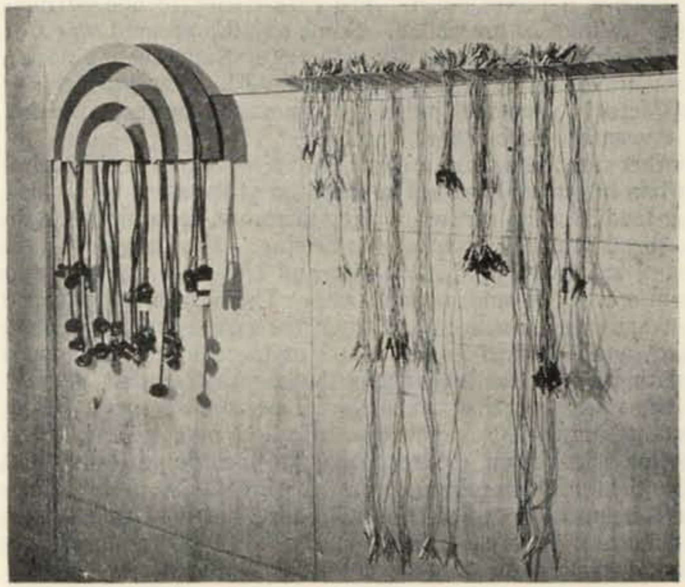


Fig 2.18—How to store leads so that the right one can be selected without loss of time and temper. (N. V. Philips Gloeilampenfabrieken)

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. 2.18 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 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 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 much 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 radio and electronics there are very many different things to be measured and tested, and there are usually several—often many—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 it 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.

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. 5.4 for a simple example of instrument inaccuracy.) The subject is discussed at length by W. H. F. Griffiths in *Wireless Engineer*, March 1943, pp. 109–126.

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 per cent. 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

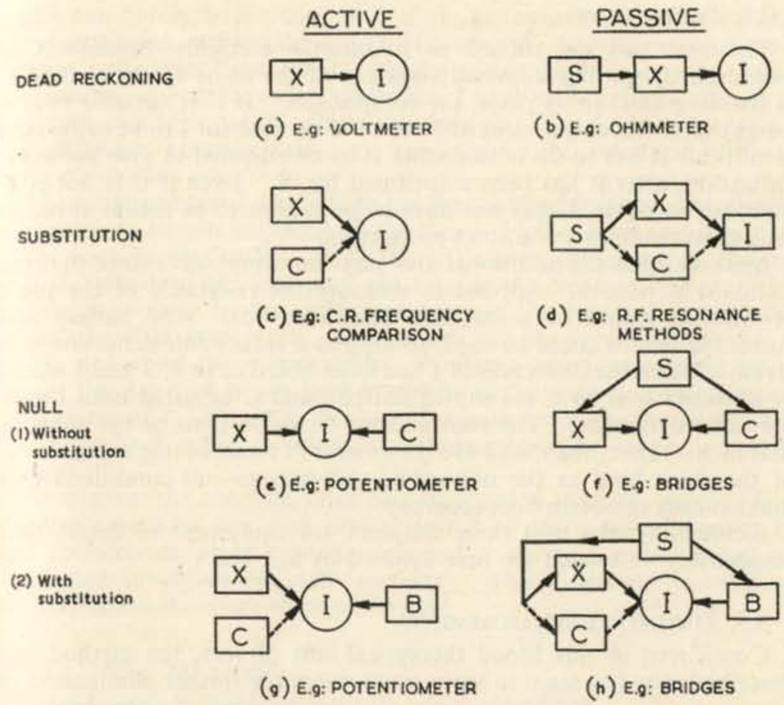


Fig. 3.1—Functional diagrams of the basic methods of measurement. X = unknown—to be measured; I = indicator; S = source of power or signal; C = standard of comparison; B = balancer

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 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 per cent, but this is of little advantage if incidental circuit resistances amount to perhaps 2 per cent 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 per cent. It can be reduced to 0·25 per cent 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. 9.28) 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

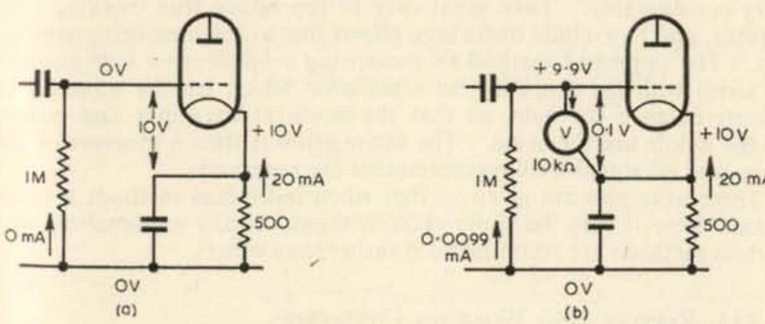
You yourself, the observer, are one of those instruments, and (with all due respect) perhaps not among the most reliable. It is extraordinarily easy to make slips in reading scales, noting the positions of range switches, connections, and conditions generally, and in performing calculations and drawing 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



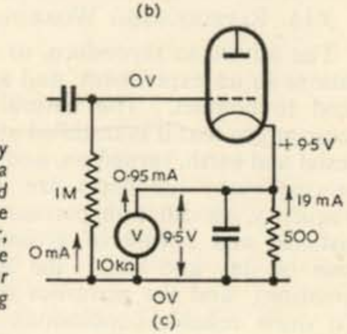


Fig. 3.2—How the measuring instrument may affect the thing measured. a shows part of a valve circuit with the conditions of current and voltage marked. If one tries to measure the voltage between cathode and grid with a voltmeter, condition b results, in which the quantity to be measured is reduced to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

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. 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 such 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 has been a broad distinction between audio and radio frequency, the division being at about 20 kc/s; but the range of vision (or "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 reach-me-downs 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.

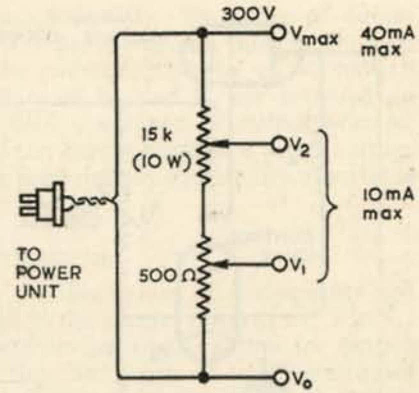


Fig. 4.1—Potential divider for extending the usefulness of a power unit—in this example, 300 V

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 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 1 V, from 1–280 V, for which constructional details of a 100-W model are given by H. E. Styles in *Wireless World*, August 1951.

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 in *Wireless World*, Sept. 1952, pp. 374–6.

4.3. Voltage Stabilizers

For many purposes it is desirable, or even essential, for the output voltage of the power unit to be stabilized against variations of input voltage and output current. Some of the means used 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. The design of the series-control d.c. type is discussed fairly fully by the author in *Wireless World*, October to December inclusive, 1948; the following is no more than an outline.

In Fig. 4.2, where for the time being R_4 should be regarded as open-circuited and R_5 short-circuited, the whole of the load current passes

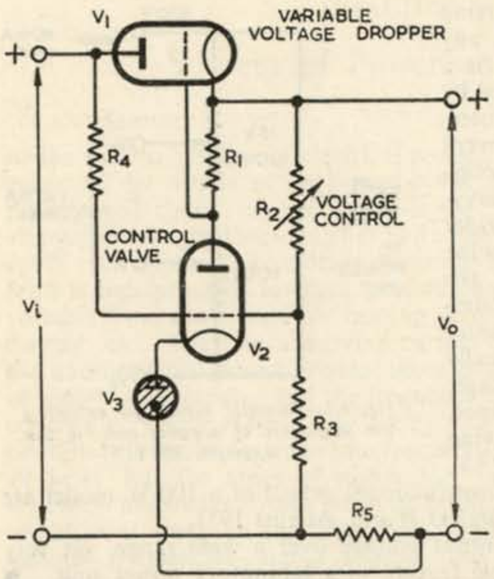


Fig. 4.2—Outline circuit of a power-supply voltage regulator for obtaining a stabilized voltage adjustable over a wide range. R_4 and R_5 are supplementary devices for improving the stabilization

through a valve V1. 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 controlled. As the difference between maximum and minimum output voltage, plus a cushioning voltage to deal with the maximum variations, plus the unavoidable drop across the valve when carrying full load current with negative grid, may be considerable, V1 must have an adequate power dissipation rating.

The voltage drop across V1 is controlled by negative grid bias derived from R_1, in the anode circuit of a control valve V2. Unwanted variations in output voltage (or at least as large a fraction of them as possible) are passed to the grid of V2, 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 neon tube V3 is a device for raising the cathode potential of V2 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 of V2,

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

thereby aiding the suppressing action in V1. 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 V2 have a similar correcting effect. The values of 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.

4.4. Design Procedure for Stabilizers

The procedure for determining suitable values of components and types of valves is explained in detail in the articles referred to. Fig. 4.3 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

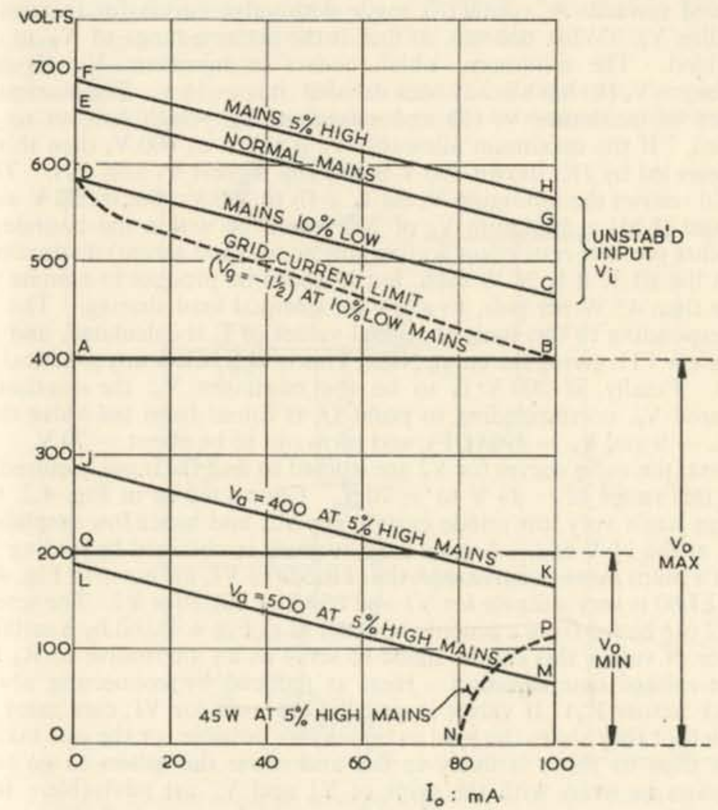


Fig. 4.3—Example of design chart for series-control voltage stabilizer

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 per cent 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 per cent 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 per cent low mains) and corresponds to full I_0 (point B). As I_0 is reduced towards A, contact is made with valve curves for increasing 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 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 V2 are studied to find the input required to give this range of $-1\frac{1}{2}$ V to -70 V. Connected as in Fig. 4.2, the system has a very low anode current for V2, and hence low amplification, at the $1\frac{1}{2}$-V end. A great improvement is obtained by feeding V2 from a point more positive than the cathode of V1, as shown in Fig. 4.4. The EF80 is very suitable for V1 and 85A2 or 90C1 for V3. The screen of V2 can be fed from a potential divider as in Fig. 4.4, 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 V1, 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. Resistors in series with the grids of V1 and V2 are advisable. It is normally necessary to provide independent heater windings on the

supply transformer for V1 and V2. Precautions must be taken to prevent V2 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. 4.6).

Some data on this type of stabilizer are given by F. A. Benson in *Electronic Engineering*, March 1952, and a design for fulfilling extremely stringent requirements where battery power had hitherto been considered indispensable is given by D. L. Johnston in *Wireless Engineer*, September and October 1947, under the title "Electro-encephalograph Amplifier".

A modified arrangement, due to E.M.I. Laboratories, in which a negative as well as a positive stabilized supply is obtained without using

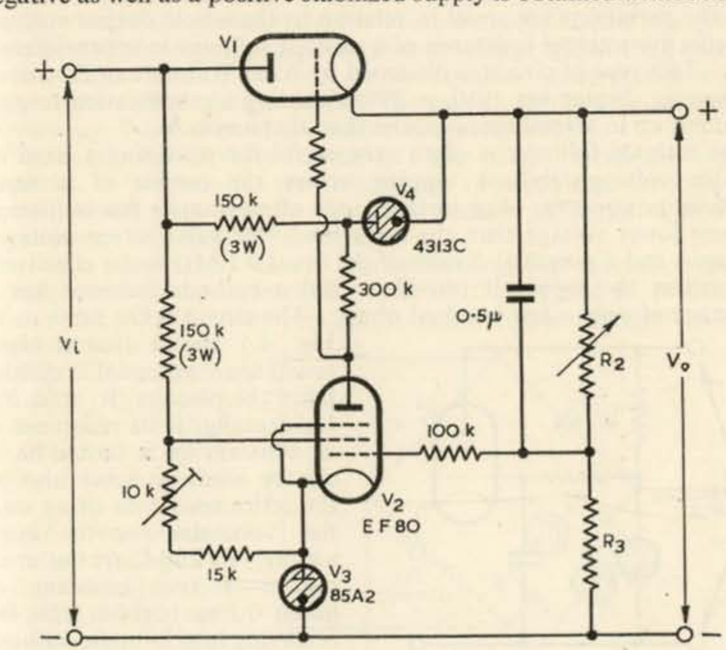


Fig. 4.4—Modified circuit for obtaining a high degree of stabilization over a wide range of voltage

any more valves, but at a sacrifice of performance and flexibility, is described in *Electronic Engineering*, June 1945, p. 559.

In the type of stabilizer described so far, a valve in series with the load absorbs surplus voltage. The shunt type, in which a valve in parallel with the load absorbs surplus current, may offer a simpler solution in some situations. Its design is dealt with by J. McG. Sowerby in *Wireless World*, June 1948.

Where a moderate degree of stabilization at a fixed voltage is acceptable, an extremely simple type of shunt stabilizer consists of a suitable type of neon tube, as listed by the valve makers for this purpose, connected in parallel with the load, the source being fed through a current-limiting resistance. The design of this type is discussed by J. W. Hughes in *Wireless Engineer*, August 1947.

4.5. Cathode-follower Stabilizers

An improvement on the direct neon-tube stabilizer, enabling a larger current to be controlled slightly more effectively, and with reduced noise voltage, is shown in Fig. 4.5. The output voltage is greater than the voltage drop across the voltage-regulator tube V_1 by the amount of grid bias needed by V_2 to pass the load current, and of course this varies with load current and supply voltage; but if a high-μ low-r_a valve is used the variations are small in relation to the whole output voltage. The effective internal resistance of a cathode follower is approximately $1/g_m$. This type of circuit is discussed by A. P. Willmore in *Electronic Engineering*, September 1950, p. 399, including a modification to give an output up to several times greater than that across V_1.

The cathode follower is often very useful for providing a fixed or variable voltage-stabilized tapping across the output of a main stabilized power unit. For instance, one often wants a few milliamps at some lower voltage than the main load—for valve-screen voltages perhaps—and a potential divider of the usual 0·1 MΩ order effectively unstabilizes the supply it provides. But a cathode follower has a resistance of only a few hundred ohms. The circuit is the same as in Fig. 4.5 except that a high-resistance potential divider takes the place of R_1 and V_1. Incidentally, if its resistance is as constant as it should be, it can be made to serve also as the series resistance of an output voltmeter for the main supply. R_2 and C are not at all critical—a time constant of about 0·1 sec is about right for removing hum or noise without making the voltage control sluggish. The design of cathode

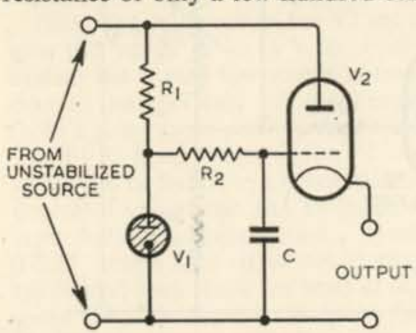


Fig. 4.5—Neon-tube and cathode-follower voltage stabilizer

followers for this purpose is discussed more fully by the author in *Wireless World*, January 1949.

4.6. 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 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 30–1,200 VA maximum, there is the Westinghouse "Stabilistor", described by A. H. B. Walker in *Wireless World*, November 1944, and very fully by the makers in their Publication EE2. 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.

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 per cent, with harmonic distortion less than 1 per cent. The design is fully described by him in *Electronic Engineering*, September to December inclusive, 1950.

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 heaters in series (*Electronic Engineering*, February 1952, p. 65).

A very simple and cheap arrangement giving several watts of constant-voltage a.c. is described by L. B. Cherry and R. F. Wild in *Proc. I.R.E.*, April 1945, pp. 262–7, and is quite suitable for such purposes as valve heating where the waveform distortion does not matter. Circuit values suitable for a 200-V supply, taken from this paper, are shown in Fig. 4.6. The variation in output voltage for an input voltage variation of ± 10 per cent is given as ± 0·35 per cent. The power efficiency is 30–50 per cent.

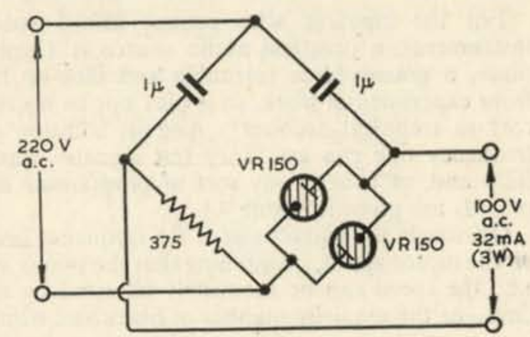


Fig. 4.6—Simple bridge stabilizer for low-power a.c.

G. N. Patchett reviews all kinds of a.c. and d.c. voltage stabilizers in *Automatic Voltage Rectifiers and Stabilizers* (Pitman, 1954).

4.7. Batteries

Although maintenance is simplified by running as much as possible off the mains, it may be difficult to do without batteries altogether. For l.t., especially if experiments demand fairly heavy current with the minimum of voltage drop, large accumulators are desirable; but there are many purposes for which the small portable type is the only practicable one. The fact that it can be brought right up to the terminals of its load is likely to result in less voltage drop than by using a large stationary battery that requires long leads to connect it to the work. It is surprising how difficult it is to maintain a full 2 V at the end of several yards of wire, even when the wire is multi-strand cable and the current is no more than an amp. Unfortunately the type of cell that gives the least variation of voltage with current drawn is not the type that stands up best to lab conditions—i.e., long idle periods and irregular charging. This must be considered when selecting batteries; the makers should be consulted. A battery should be permanently installed in some place where it will not be in the way, but on the other hand not so much out of the way that it is too much bother to inspect it fairly frequently for acid level, creepage, or corroded terminals.

Although it does not give such a constant voltage as the acid type, the Venner alkali accumulator* is a fraction of the weight and can be allowed to remain discharged indefinitely; it gives 1·5 V, suitable for 1·4-V valves, and is particularly useful for small portable equipment.

At one time a h.t. accumulator was almost essential to a radio laboratory, but it was expensive and troublesome, and nowadays a stabilized mains unit can be designed for even the most exacting requirements. Nobody with experience of h.t. accumulators is likely to regret this development.

4.8. Signal Sources: The Gramophone

For the amateur who cannot afford specialized and expensive instruments, a practical audio source is a gramophone. In the first place, a gramophone turntable and pick-up has its uses quite apart from experimental work, so it may not be necessary to charge all of its cost to technical account! And in addition to pure tones of fixed frequency one 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 Table 4.1.

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

* Venner Accumulators, Ltd., Kingston By-pass, New Malden, Surrey.

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 80 r.p.m. it is 75. In general, it is

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

But one must beware of abnormal supply frequency during periods of overloading.

As for the pick-up, the requirements are identical with those for first-class gramophone reproduction—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. The piezo-electric crystal type yields a large output voltage, but its high and capacitive impedance demands special care in circuit arrangement. 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

Table 4.1

Supplier	*Number*	*Side*	*Recording*	*Reference level*
British Sound Recording Association	PR.103 (vinyl disk) or PR.103/S (shellac disk)	1	Constant tones: 10, 9, 8, 7, 6, 5, 4, 3, 2, 1 kc/s (0 db); 500 c/s (−1 db); 200 c/s (−4·5 db); 100 c/s (−10·5 db); 50 c/s (−17·5 db); recorded in bands as continuous spiral.	2·25 cm/sec (1 cm/sec + 7 db)
		2	Ditto	
Decca	K.1802	1	Gliding tone: 14–3 kc/s	3·25 cm/sec (1 cm/sec +10·25 db)
		2	Gliding tone: 3 kc/s–10 c/s	
	K.1803	1	Gliding tone: 14–3 kc/s	
		2	Gliding tone: 3 kc/s–10 c/s	
			ffrr characteristics, with rising velocity above 3 kc/s. 0 db above 300 c/s; −1 db at 250 c/s; −5 db at 100 c/s; −11 db at 50 c/s.	
	K.1804	1	Constant tones: 14, 13, 12, 11, 10, 9, 8, 7, 6, 5 kc/s (0 db).	
		2	4, 3, 2, 1 kc/s and 400 c/s (0 db); 250 c/s (−2 db); 100 c/s (−6 db); 55 c/s (−10·5 db); 30 c/s (−16·5 db).	

All 12-in records, for 78 r.p.m.

[*Table continued overleaf*

Table 4.1 (*continued*)

Supplier	*Number*	*Side*	*Recording*	*Reference level*
E.M.I. Studios	JG.449	1	Constant tones: 20, 18, 16, 14, 12, 10, 8, 6, 4·5, 3·5, 2 kc/s (0 db); 500 c/s (−1 db); 160 c/s (−5·5 db); 70 c/s (−12 db).	1 cm/sec
		2	19, 17, 15, 13, 11, 9, 7, 5, 4, 3, 1 kc/s (0 db); 250 c/s (−3 db); 100 c/s (−8·5 db); 50 c/s (−14 db).	
	JH.138	1	Eleven bands of 60 c/s and 2 kc/s added; first band at 8·6 db and 10·3 db respectively; each subsequent band 2 db below previous one.	
		2	Eleven bands of 400 c/s and 4 kc/s added; first band at 22·5 db and 10·5 db respectively; each subsequent band 2 db below previous one.	
H.M.V.	DB.4033	1 & 2	Special sound demonstrations.	
	DB.4034	1	Constant tones: 8·5, 8, 7·5, 7, 6·5, 6, 5·5, 5, 4·5 kc/s.	
		2	4,000, 3,750, 3,500, 3,250, 3,000, 2,750, 2,500, 2,250 c/s.	
	DB.4035	1	2,000 1,800 1,600 1,400 1,200, 1,100, 1,000, 900 c/s.	
		2	850, 800, 750, 700, 650, 600, 550, 500 c/s.	
	DB.4036	1	450, 425, 400, 375, 350, 325, 300, 275 c/s.	*
		2	250, 225, 200, 180, 160, 140, 120, 100 c/s.	
	DB.4037	1	90, 80, 70, 60, 50, 40, 30, 25 c/s.	
		2	Gliding tone: 8,500–25 c/s. Level 0 db 8,500–300 c/s to −14·5 db at 25 c/s.	

All 12-in records, for 78 r.p.m.

" Level " refers to r.m.s. lateral velocity of groove, to which the electrical output of a perfect pick-up is proportional.

* Levels specified by makers: nearly constant 8,500–250 c/s; −11·1 db at 25 c/s.

of time taken over adjustments can be saved if the output is substantially uniform over the whole band of frequencies. Whatever point is used it is most important that it be suited to the record in use. Normally it will be a more or less permanent stylus; on any other occasion it must be remembered that the output amplitude is liable to vary with different needles even of the same type. The needle should therefore be used for as short a time as possible—on

the higher frequency grooves at any rate. The top frequency calls for a brand-new needle, so it is customary to begin tests at that end.

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. The 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 can be very helpful for taking down readings during experiments, especially those requiring much concentration and speed.

4.9. Valve Oscillators: General Requirements

All-electric signal sources consist of valve oscillators in more or less elaborated forms. In dealing with oscillators 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. 4.11 to 4.16, the function of the valve or valves 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 (or can be considered as being) in parallel with the tuning circuit, so the positive resistance 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 the valve characteristics, such as grid current that increases the positive conductance, 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 a lot of surplus has to be provided, in order to maintain oscillation over a large range of positive conductance, 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 though this effect is small it is important.

The three main characteristics—frequency, waveform and amplitude—are thus closely bound up together, so in Sec. 4.17 special attention will be devoted to the means by which amplitude is limited.

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

4.10. 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 simple parallel tuned circuit (Fig. 4.7) is fed with currents of various frequencies. The ingoing current of the fundamental frequency

Table 4.2

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

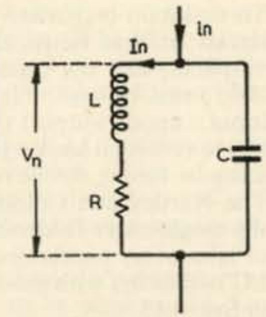


Fig. 4.7—Here i_n, I_n and V_n denote currents and voltage at the nth harmonic of the frequency to which the circuit is tuned

(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 4.2, 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 occur in practice. If the maintaining current had, say, 30 per cent second harmonic, the voltage across LC would have $30/150 = 0{\cdot}2$ per cent 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 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.

4.11. Feedback LC Oscillators

The first and largest class of valve oscillators comprises those that consist of an amplifier with frequency-discriminating positive feedback.

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

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 amplifier (grid input; anode 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 of the single-valve feedback LC 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 valve in the amplifier are considered in Sec. 4.15.

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. 4.8 shows an example of a highly frequency-stable type of oscillator embodying these ideas. It is sometimes called the Clapp oscillator, because it was described by J. K. Clapp in *Proc. I.R.E.*, March 1948; but it appears to be first due to G. G. Gouriet, who has written about it in *Wireless Engineer*, April 1950, pp. 105–112. 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 leak and grid capacitor, L_1 is a r.f. choke to provide a conductive path to the cathode and C_4 a large capacitance to hold the anode at zero signal potential. The valve can of course be a tetrode or pentode.

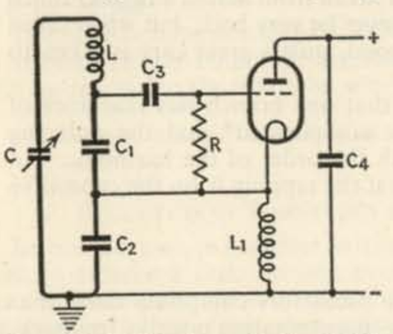


Fig. 4.8—Modification of the Colpitts oscillator circuit to give exceptional stability of frequency

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.

Guiding principles in designing positive-feedback LC 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.

Because L and C in Fig. 4.8 are the main frequency-determining elements, 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_1C_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" (*Wireless World*, August and September 1952).

4.12. Electron-coupled Oscillators

The Fig. 4.8 circuit is a very good example of avoiding undesired frequency variations due to the maintaining valve. But this precaution would be more than nullified if the oscillator were connected up as a signal source in such a way as to affect the tuning; e.g., by taking the signal direct from the tuning circuit. One 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. 4.8, the cathode is not kept at constant potential but is tapped up the oscillatory circuit, though more usually on the coil rather than the capacitor side of it, as in Fig. 4.9. But this feature is no essential part of electron-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 (g_3) serves as a screen, and one might as well 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.*, December 1931, pp. 2095–2108) 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, as in Fig. 4.9, it is possible to balance out the effect of varying supply voltage on frequency. It may therefore be considered worthwhile using a tetrode to gain this advantage, which is not obtainable

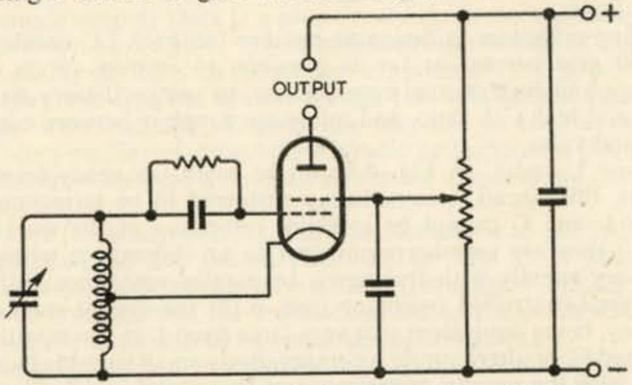


Fig. 4.9—Electron-coupled tetrode circuit for stabilizing the frequency of oscillation against supply-voltage variations

with a pentode. If so, the device in Fig. 4.9 is more convenient than neutralization. But 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.

4.13. 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 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 *Wireless Engineer*, October 1933, pp. 527–540.

The action of the dynatron depends on secondary emission from g_2. The curves of the Mazda AC/S2 in Fig. 4.10 show that when the anode voltage V_a is between the limits of about 10–90 per cent 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. 4.11 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

* "Electron-coupled Oscillators", *Wireless World*, December 1952, pp. 515–8.

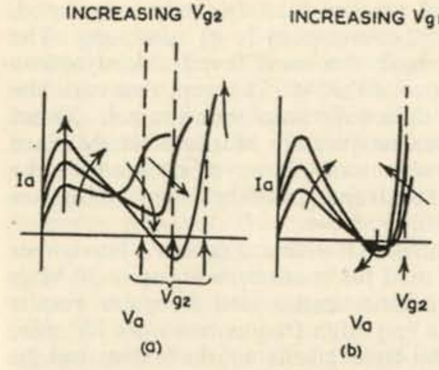


Fig. 4.10—Typical anode-current/anode-voltage curves of dynatron, showing the effects of varying g_2 voltage (a) and g_1 voltage (b). V_{g2} is marked along the anode voltage scale

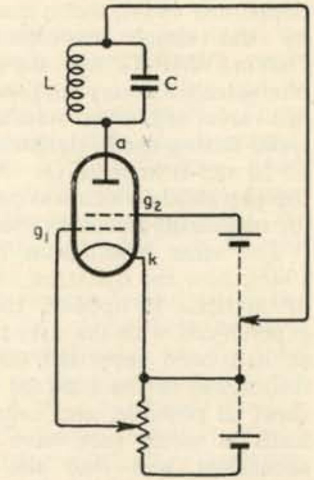


Fig. 4.11—Basic dynatron circuit

sufficient amplitude of oscillation, and to use V_{g1} for reducing the conductance until oscillation is only just maintained. Under these conditions oscillation sweeps 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 the AC/S2 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 low capacitance and damping of the terminal to which the high-potential end of LC is connected; 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. 4.18), 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 (*Wireless World*, September 1944), that a number of types of pentode can be substituted, by the simple expedient of "commoning" g_2 and g_3. The Osram VMP4G was shown to have the most favourable dynatron characteristics, very similar to those of the AC/S2 except that variable-mu valves are not so suitable for automatic amplitude control. Other types having useful dynatron characteristics are Mazda SP41, Mullard EF50 and Brimar 6J7G. Modern pentodes, however, do not have the top-cap anode which was one of the advantages of the original dynatron for measuring purposes (Sec. 9.30).

The same investigator has shown (*Wireless Engineer*, November 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.

4.14. 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.*, October 1935) but so called by Brunetti, whose paper on it in *Proc. I.R.E.*, December 1937, is worth studying also for its explanation of "average negative resistance". LC is connected

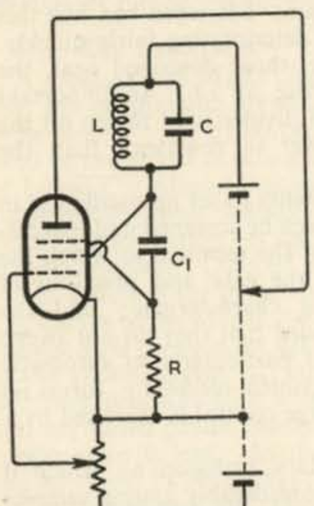


Fig. 4.12—Basic transitron circuit

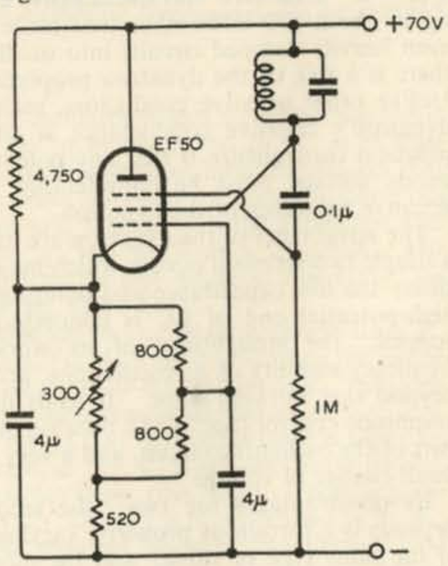


Fig. 4.13—Practical transitron circuit

in the g_2 circuit (Fig. 4.12) 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. 9.31). 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 (*Wireless World*, March and April 1943) and F. P. Williams (*Wireless World*, August 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. 4.13.

What might be called an electron-coupled transitron, in which the various grids in a 6A8G heptode are used to generate r.f. and a.f. and the modulated output is taken from the anode, has been devised by K. W. Mitchell in an inexpensive signal generator described by him in *Wireless World*, February 1945, pp. 56–7. As usually happens when a single valve is made to do a number of different things, conditions cannot be optimum for all at once. And there appears to be no alternative to the type of valve specified. But within its limits this oscillator has been found very useful.

4.15. Amplifier Two-terminal LC Oscillators

The need for more than two connections to the tuned circuit in the feedback oscillators referred to in Sec. 4.11 is to obtain the phase reversal to cancel the reversal in the single-valve earthed-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. 4.14, is

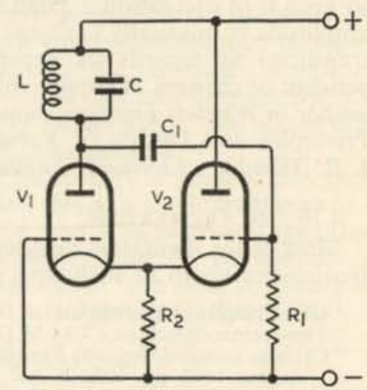


Fig. 4.14—Outline circuit of cathode-coupled oscillator

described by F. Butler in *Wireless Engineer*, November 1944. It can be regarded as developing a negative conductance between the LC terminals. V1 is an earthed-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 V2. 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.

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., in *Proc. I.R.E.*, June 1946, pp. 502–5.*

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 across R_2 may be very distorted (Sec. 4.10).

The greatest negative conductance is obtained at a value of R_2 which, with usual valves, is of the order of 1 kΩ.

The performance of what J. R. Tillman classifies as a precision negative resistance is detailed by him in *Wireless Engineer*, December 1947, pp. 357–371. It 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 in *Wireless Engineer*, January 1945, pp. 17–24. (See also " The Principles and Design of Valve Oscillators ", by A. C. Lynch and J. R. Tillman. *Electronic Engineering*, February and March 1945.)

4.16. RC Oscillators

Most 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

* Other helpful references are:
" Two-terminal Oscillator ", by M. G. Crosby. *Electronics*, May 1946, pp. 136–7.
" Cathode-coupled Negative Resistance Circuit ", by P. G. Sulzer. *Proc. I.R.E.*, August 1948, pp. 1034–9.
" Frequency and Amplitude Stability of the Cathode-coupled Oscillator ", by P. G. Sulzer. *Proc. I.R.E.*, May 1950, pp. 540–2.

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.

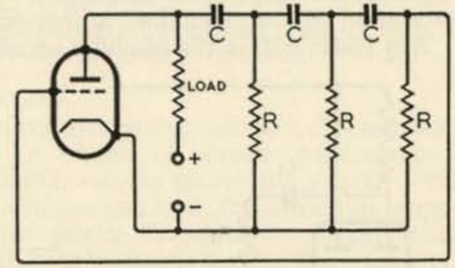


Fig. 4.15—Outline circuit of a simple form of phase-shift oscillator. The frequency of oscillation depends on the values of R and C

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. 4.15.

The frequency at which the phase shift is 180°, and therefore—at least approximately—the frequency of oscillation is

$$f_0 = \frac{1}{2\pi CR\sqrt{6}}$$ (see reference to Vaughan on p. 67 (top))

and the voltage amplification needed to make good the loss in the three-stage RC 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 $f_0 = (\sqrt{6})/2\pi CR$, the choice usually depending on which is used for frequency control. Frequency can be varied over a very wide range by using three ganged capacitors and step-switched resistors, or vice versa.

Since the RC 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 provided the gain of the valve is adjusted so as barely to offset the loss through the RC network a very good sine waveform can be obtained. To ensure this, automatic control (Sec. 4.17) is almost indispensable.

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. 4.16, which is the RC counterpart of the cathode-coupled LC oscillator (Fig. 4.14). The only frequency at which the phase shift in R_2C_2 is equal and opposite to that in R_1C_1 is

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

* " The Equivalent Q of RC Networks ", by D. A. H. Brown. *Electronic Engineering*, July 1953, pp. 294–8, and September 1953, pp. 394–5.

and the required voltage gain is $1 + R_1/R_2 + C_2/C_1$. When, 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 = $9\frac{1}{2}$ db). If therefore the gain of the amplifier is $9\frac{1}{2}$ 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

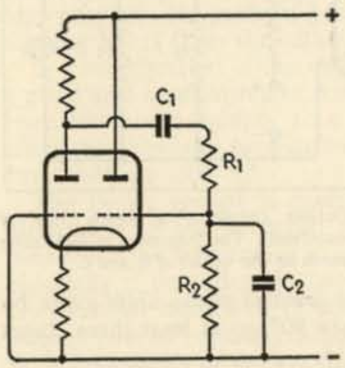


Fig. 4.16—A particularly simple type of RC oscillator

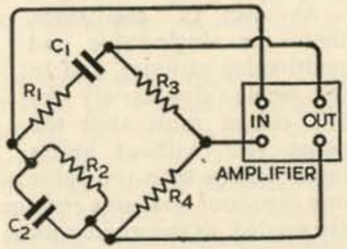


Fig. 4.17—Functional diagram of bridge-controlled RC oscillator

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. 4.17).

Another way of looking at the circuit is to consider it as a bridge (actually known as the Wien bridge) which would be balanced if $R_3 = 2R_4$, because both of the " detector " points would be at the 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 $9\frac{1}{2}$ 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 a practical example is described in Sec. 4.21.

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*, February 1948, pp. 178–186) H. A. Whale has shown that the greater C_2/C_1 and R_1/R_2 the better the frequency stability.

RC oscillators in general were reviewed by J. A. B. Davidson in *Electronic Engineering*, January and February 1944; there is an elementary treatment in *Wireless World*, September 1950, pp. 331–4; a full treatment of the 180°-shift type by W. C. Vaughan (*Wireless Engineer*, December 1949, pp. 391–9); a constructional description of a B.B.C. oscillator of the type shown in Fig. 4.17, by A. R. A. Rendall and F. A. Peachey (*Wireless Engineer*, February 1948, pp. 37–43); and an article on a type of RC oscillator that differs somewhat from both the types described here, with the object of convenient frequency control, by W. G. Raistrick (*Wireless World*, November 1950, pp. 409–411).

4.17. Amplitude Stabilization

As explained in Sec. 4.9, the key to desirable oscillator characteristics is the 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 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 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 of the best is the resistance control shown in Fig. 4.18. The circuit is so designed that when R is adjusted until oscillations just start it is as large as possible and at least several times r_a. The valve is worked Class A, and when R is properly adjusted a mere trace of grid current is sufficient to stabilize oscillation. Unfortunately, the method is restricted to a few hundred kc/s; at higher frequencies, stray capacitance across R tends to by-pass it. For the design of this type of circuit see F. E. Terman's *Measurements in Radio Engineering* (McGraw-Hill), pp. 283–7.

The use of negative feedback in Tillman's oscillator (mentioned in Sec. 4.15) 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. 4.10).

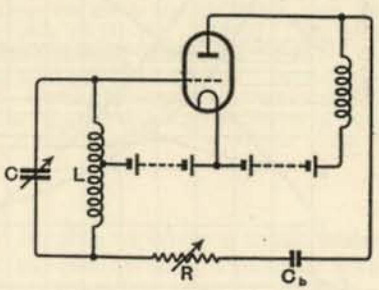


Fig. 4.18—Resistance-stabilized oscillator circuit suitable for a.f. and low r.f.

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 in *Wireless World*, June 1947, p. 219.

4.18. Automatic Amplitude Control

To avoid the inconvenience and instability of manual control, several automatic systems have been devised. In a sense every oscillator has automatic control, by the mechanism described in Sec. 4.9, 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. 4.8 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.*

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

* See L. B. Arguimbau in *Proc. I.R.E.*, January 1933, pp. 14–28.

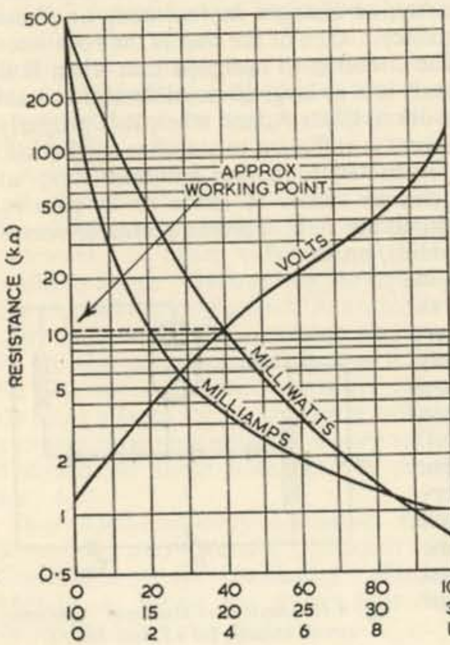


Fig. 4.19—Characteristics of Standard Telephones thermistor, Type A55/13/100. A suitable working point is marked. For an example of its use, see Fig. 4.22

"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. If the resistance of R in Fig. 4.18, for example, increased considerably with temperature, and if its initial value ensured oscillation at all settings of C, then directly oscillation started there would be a current through R which would raise its resistance and so reduce feedback, preventing further increase in amplitude. Ordinary metal-filament lamps behave in this way, but the values of resistance obtainable are too low for this type of circuit, though they have been fairly widely used for controlling RC oscillators; e.g., R_4 in Fig. 4.17.

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 the opposite of what is required in Fig. 4.18, but is eminently suitable for R_3 in Fig. 4.17. A practical example is given in Sec. 4.21.

Fig. 4.19 shows the characteristics of one type of thermistor—the Standard Telephones A55/13/100. 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. For example, a suitable working point is marked in Fig. 4.19, just over 10 kΩ, corresponding to a dissipation of 40 mW (20 V, 4 mA).

In a bridge or other system for magnifying the rate of control, such as that in Fig. 4.17, 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.*, January 1948, pp. 16–19). 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.*, October 1938, pp. 1278–95; see also "Variable-frequency Bridge-type Frequency-stabilized Oscillators" by W. G. Shepherd and R. O. Wire (*Proc. I.R.E.*, June 1943, pp. 256–269), and "Thermistor-regulated Low-frequency Oscillator" by L. Fleming (*Electronics*, October 1946, pp. 97–100). The design of a lamp-in-bridge controlled 20-kc/s oscillator of very high frequency stability is detailed by T. Roddam in *Wireless World*, August 1948, pp. 286–8.

4.19. The Beat-frequency Source

For some purposes it is helpful to be able to cover a much wider range of frequency with one sweep of the control than is possible with any of the foregoing 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 range of 10 : 1 is easily attainable. But the usually accepted limits of the audio range are 20–20,000 c/s or 1,000 : 1; and the modulation frequencies in television are much wider still—from zero up to several megacycles per second.

An unusual type of RC oscillator covering from 20 c/s to 3 Mc/s in one range is described by F. B. Anderson in *Proc. I.R.E.*, August 1951, pp. 881–890. But for 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. A treatise on the design of the b.f. source is outside the scope of this book, but it may be just as well to indicate some of the requirements, if only to deter

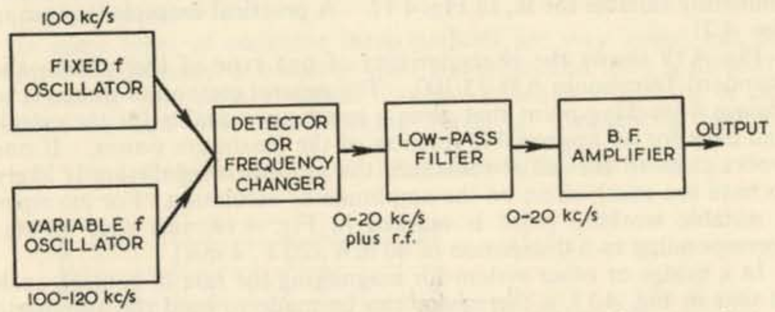


Fig. 4.20—Block diagram of beat-frequency a.f. source, with typical frequency values

thrifty readers from too light-heartedly setting out to make one. The essential stages are shown in Fig. 4.20, 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 the important business of 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 strong harmonics, to which the apparatus may be much more responsive, the result does not indicate the true 50-cycle response at all.

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 ", *Wireless Engineer*, 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 is generally shaped to give a logarithmic scale between 100 and 10,000 c/s, and linear outside these limits.

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. Triodes are far less dependent on accurate load matching than are other types, but one of the advantages of negative feedback is that it can be used to make any valve have a very low output impedance, so that load matching is not critical. 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 per cent 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.

An advanced design is detailed by C. G. Mayo and D. G. Beadle in *Electronic Engineering*, October 1951, pp. 368–373.

4.20. Commercial A.F. and V.F. Sources

The Sullivan-Ryall b.f. source deserves mention because of the amazing standard achieved—harmonic content 0·1 per cent over the telephonic frequency range and less than 0·3 per cent over the whole range at 0·5 W; maximum output 5 W; output voltage constant to ±0·1 db; frequency stability within 0·2 per cent ±1 c/s. Valuable articles by W. H. F. Griffiths on the design of b.f. oscillators, including a description of the Sullivan-Ryall type, are in *Wireless Engineer*, May 1934 and July 1935.

The tendency, however, is for the RC oscillator to oust the b.f. type, at least for audio frequencies. A usual choice of frequency coverage

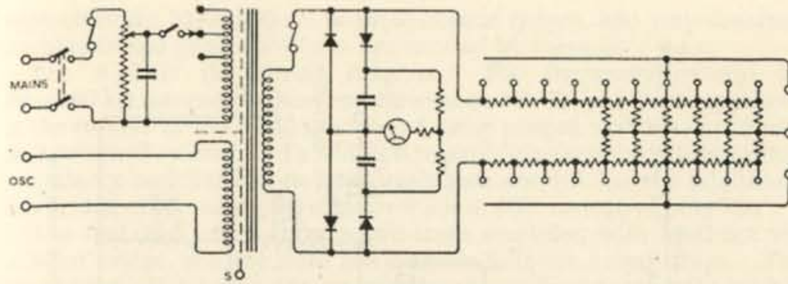


Fig. 4.21—Circuit diagram of Type LS158 Millivolter—a balanced attenuator with metered input from oscillator or mains, for providing known peak-to-peak voltages. (Allied Electronics Ltd.)

is 20–20,000 c/s, in three decade ranges; but some have ranges up to 100 or 200 kc/s. In some instruments, such as the Dawe Type 400A (described in *Electronic Engineering*, August 1947, pp. 246–9) and the Advance Type H1, the continuously-variable frequency control consists of a gang capacitor, and the ranges are obtained by switched resistors. This has the advantage of very smooth control of frequency, but the unavoidably low maximum capacitance leads to very high circuit impedances—of the order of 10 MΩ on the lowest-frequency range—and the triple capacitor and other parts have to be very carefully screened in order to exclude hum. A lower and less widely varying circuit impedance can be obtained by using the resistance elements as the frequency control, and switched capacitors for the ranges. The only objection is that with the usual type of rheostat the resistance varies in small steps, so that it may not always be possible to set the control precisely to a given frequency; but this difficulty is hardly apparent if a suitable component has been fitted.

In the Muirhead decade oscillator, a comprehensive account of which appears in *Muirhead Technique*, July and October 1947, the frequency is controlled by four 11-way switches, in steps of 1 c/s from 1–11,110 c/s, and there is a ×10 frequency multiplier.

For v.f. testing, the General Radio b.f. oscillator covers 50 c/s to 5 Mc/s in two sweeps. And for a.f. intermodulation tests (Sec. 11.16) the same firm supplies a two-signal audio generator.

The Millivolter LS158 by Allied Electronics is, in effect, a variable attenuator with input transformer and peak-to-peak voltmeter (Fig. 4.21) by which an a.f. oscillator with uncontrolled output can be converted into an a.f. signal generator giving known output voltage over the range 100 μV to 100 V; alternatively the a.c. mains can be used in place of the oscillator. Its primary purpose is for continuous voltage calibration of an oscilloscope (Sec. 5.30).

4.21. A General-purpose A.F. Source

For the experimenter who wants an easily made yet satisfactory a.f. source the following is a practical design. It is a resistance-tuned

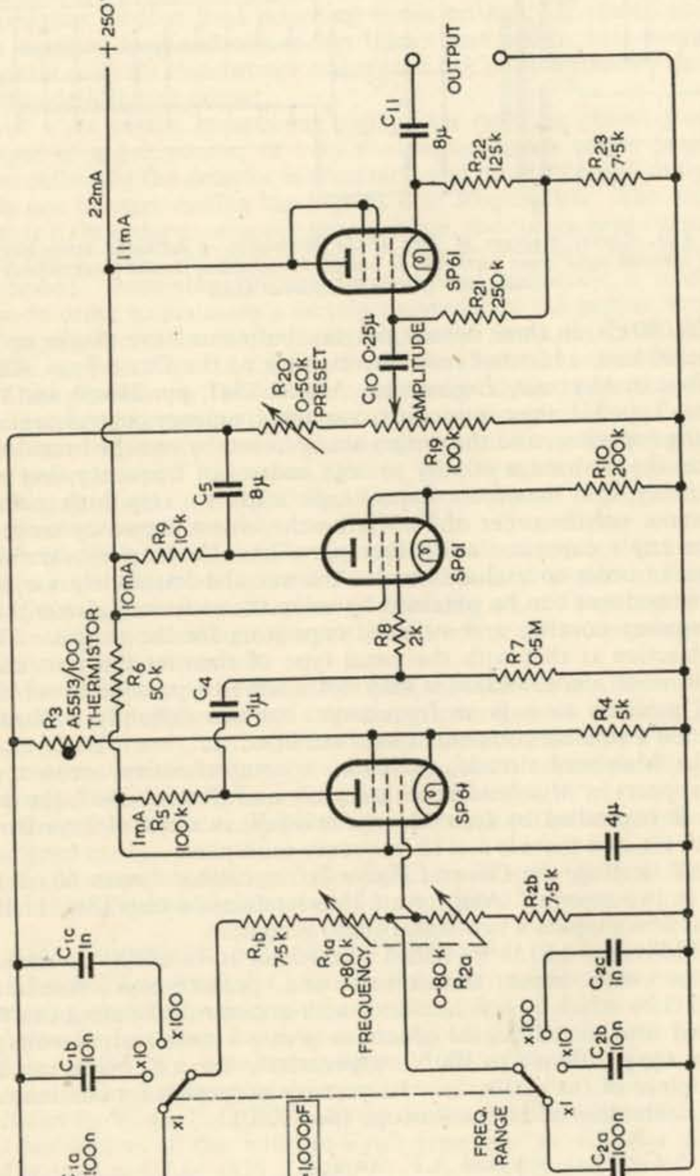


Fig. 4.22—Circuit diagram of 20–20,000 c/s audio source with amplitude stabilization and pure waveform

type covering 20–20,000 c/s in three decade ranges, and very constant amplitude and good waveform are secured by thermistor a.a.c.

Fig. 4.22 is the circuit diagram. The frequency control by 80 + 80 kΩ ganged rheostat and three pairs of switched fixed capacitors is the reverse of the usual practice of using ganged variable capacitors and switched resistors. In addition to avoiding excessively high circuit impedance and liability to hum, resistance control has the additional advantage of spreading the scale over about 300° instead of only 180°.

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 thermistor (R_3) keeps the amplitude of oscillation 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 distortion of less than 1 per cent.

A very close approximation to a logarithmic frequency scale is obtainable by using rheostats with graded 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 multiples of one another it is possible to make one scale serve all three ranges, using $\times 10$ and $\times 100$ factors for ranges 2 and 3.

The 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 200–2,000 c/s scale with approximately equal margins above and below, and then calibrate fully over that frequency 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. 4.22 is the nucleus, appears in *Wireless World*, August and September 1949. The additional features include a sine-to-square-wave converter, attenuators totalling 120 db, a second output stage with phase inverter to give a balanced output, a monitor valve voltmeter of the type shown in Fig. 5.23 (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.

4.22. 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. 9.25). 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.

4.23. 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.

4.24. 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. Fig. 4.23 shows the precautions adopted in a comparatively simple type of instrument. 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

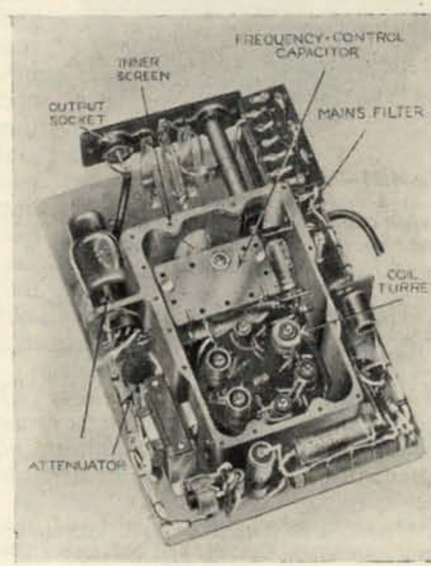


Fig. 4.23—Opened-up rear view of Avo r.f. signal generator, showing means for preventing r.f. leakage. The inner screen is a cast aluminium box with numerous fixing screws to ensure a tight-fitting cover. (Automatic Coil Winder & Electrical Equipment Co., Ltd.)

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.

4.25. 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 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.* 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 1–10,000 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, but Fig. 4.24 shows the construction of an 80-db resistive ladder attenuator, variable in four steps, and reasonably accurate up to 300 Mc/s owing to the use of carbon resistors in an aluminium die casting shaped to give good screening and to match 75 Ω impedance. For the highest frequencies it is generally necessary to use a reactive attenuator rather than a resistive one (Sec. 6.22).

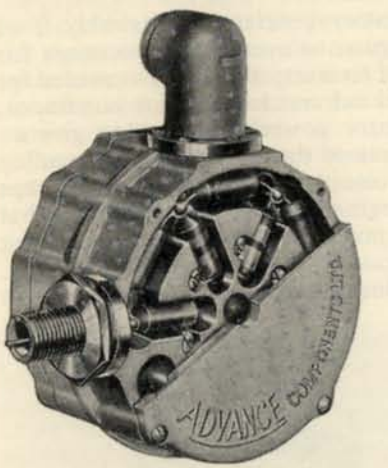


Fig. 4.24—75-Ω r.f. ladder attenuator with a maximum attenuation of 80 db in four steps. Note construction to enable use up to 300 Mc/s. (Advance Components Ltd.)

The accuracy of a microvolt calibration, which may be within 5 or 10 per cent 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

* See " Some Aspects of R.F. Attenuators ", by R. H. Mapplebeck. *Marconi Instrumentation*, October and November 1947, pp. 52–5.

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, 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. The thermionic diode, responsive to peak values, is subject to error if the waveform is non-sinusoidal, and the " zero " current—varying with heater voltage—is a difficulty, so this type has been largely superseded by the crystal rectifier, which has no zero current.

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 in *Marconi Instrumentation*, March 1954, pp. 105–112.

4.26. 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

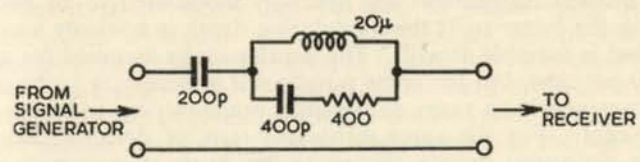


Fig. 4.25—Composition of standard dummy aerial

on receivers, a compact dummy aerial, made up as in Fig. 4.25, is inserted between the generator and the receiver under test. This composite impedance, shown in Fig. 11.20, 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

* " A Method of Calibrating Standard-signal Generators and R.F. Attenuators ", by G. F. Gainsborough. *J.I.E.E.*, Pt. III, May 1947, pp. 203–210.

so, this may lead to false comparisons when a low-impedance load is being fed, unless compensation is made by external resistance.

4.27. 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 at least one model 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.

4.28. 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 per cent, 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 per cent. 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. 11.17).

Another thing that is important to avoid 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 per cent. Anode and even screen-grid modulation require appreciable power. Control-grid modulation is perhaps the most likely to give trouble. 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 rather 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.

In signal generators for television, modulation is the main issue, and a more or less elaborate 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.

4.29. "Wobbulation"

This is a specialized type of frequency modulation, used in conjunction with an oscilloscope for observing frequency responses (Sec. 5.29). 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, and a contact mounted on the spindle is used to synchronize a linear time-base circuit (Fig. 5.36). 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, adopted in the Marconi Instruments TF.923 Television Sweep Generator,* 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 sweep can be continuously varied down to zero by controlling the moving-coil current.

Most often, however, an electronic method is used. The usual 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

* See W. F. D. Busby in *Marconi Instrumentation*, May 1952, p. 72.

C and R in Fig. 4.26. 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.

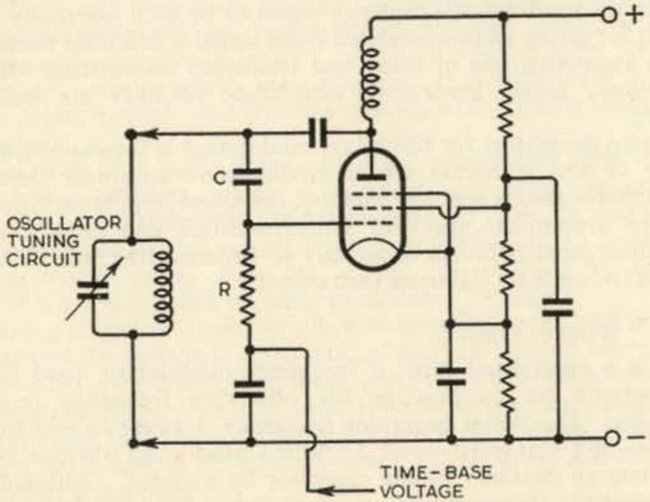


Fig. 4.26—Reactance-valve circuit, for causing the voltage of a time-base to produce corresponding frequency variations in the signal source

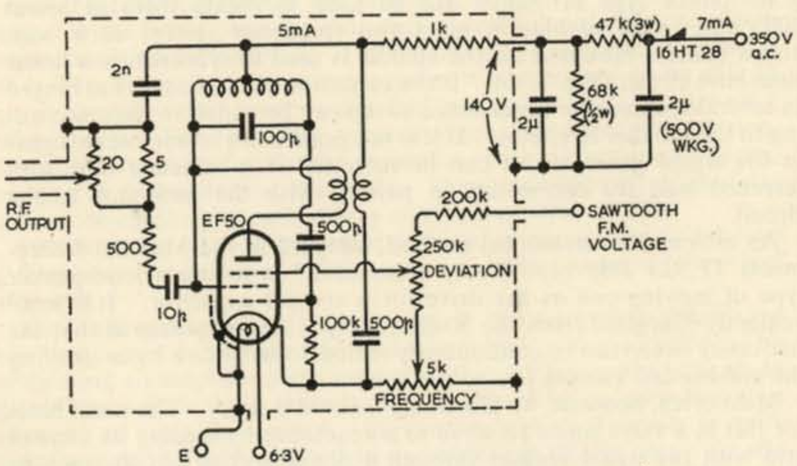


Fig. 4.27—Author's practical version of Johnson one-valve f.m. oscillator, or "wobbulator"

A similar result is obtained in a different way by a method due to K. C. Johnson, and described by him in *Wireless World*, April and May 1949. Advantages are that as much as 30 per cent 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 is given in *Wireless World*, October 1950, and is obtainable from the publisher in leaflet form. The circuit diagram is shown here as Fig. 4.27.

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. 4.19), wobbulation being applied to what would otherwise be the fixed-frequency oscillator.

4.30. 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 Advance Type E2, covering 100 kc/s—100 Mc/s in six bands of fundamental frequency. Output is 1 µV to 100 mV continuously variable, with an additional fixed level of 1 V.

In the strictly lab. class, 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 7 kc/s—and so including most of the v.f. as well as communication and sound broadcasting bands—is the Marconi TF.867. Fig. 4.28 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 per cent 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. 4.29 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 crystal calibrator. In such an elaborate system the range switch is required to change over a large number of contacts, and Fig. 4.30 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. 4.31, 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*, obtainable from Marconi Instruments, Ltd. In *Proc. I.R.E.*, October 1949, A. V. Haeff, T. E. Hanley and C. B. Smith deal with developments in the design of generators for 90–9,700 Mc/s.

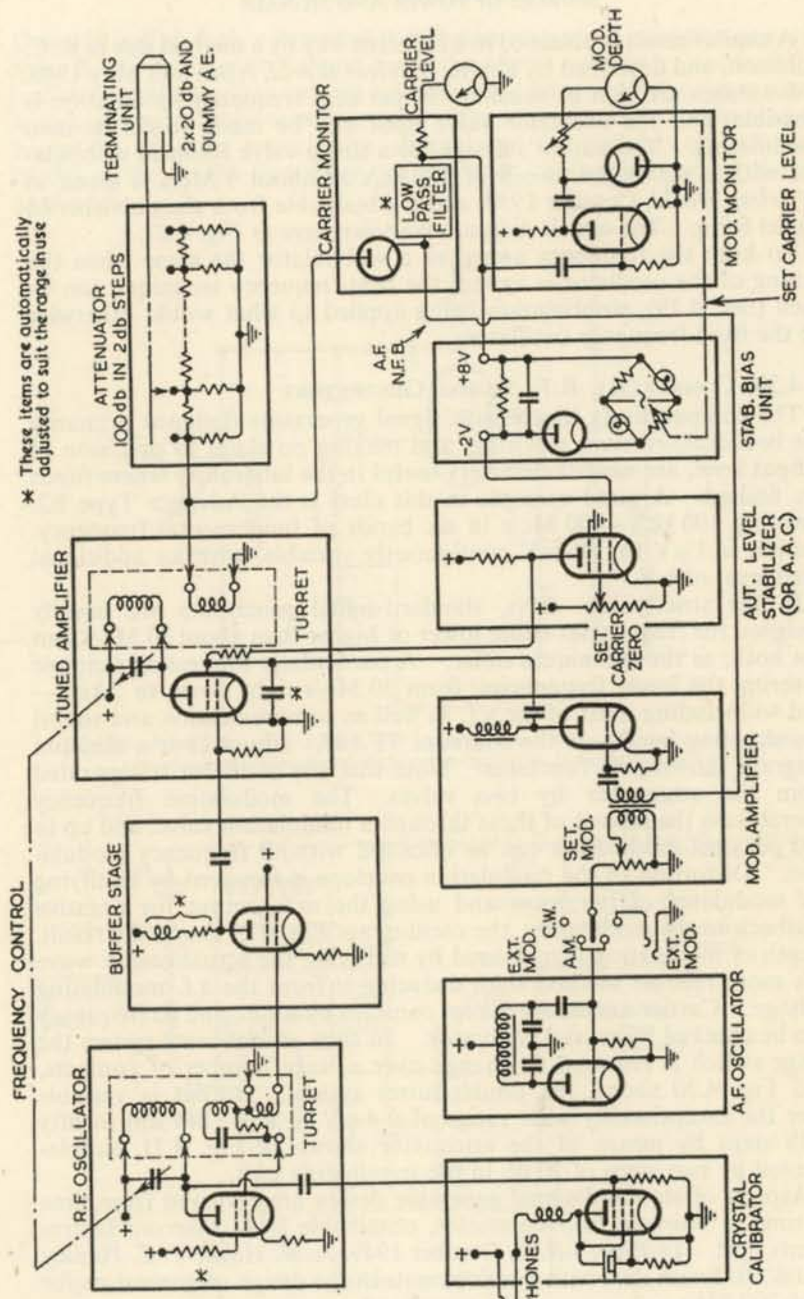


Fig. 4.28—Functional diagram of a laboratory standard-signal generator—Marconi Instruments TF.867

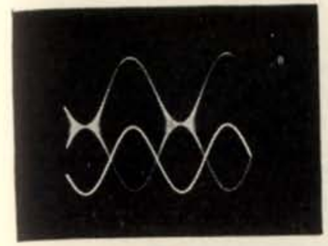


Fig. 4.29—Oscillogram of modulation envelope with modulating waveform superimposed, showing fidelity of modulation even at practically 100 per cent depth. (Marconi Instruments Ltd.)

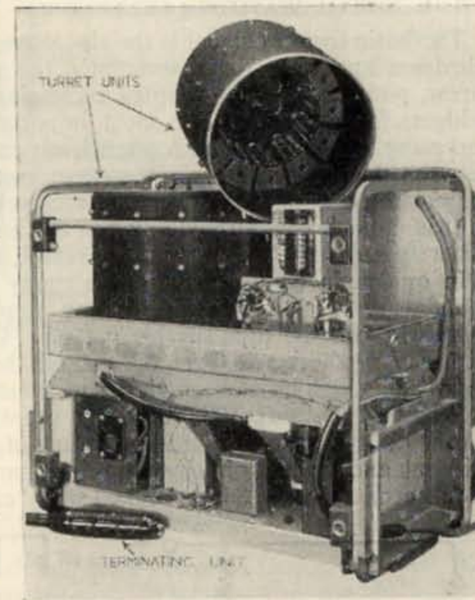


Fig. 4.30—Opened-up rear view of TF.867 standard-signal generator, with one coil turret lifted off. (Marconi Instruments Ltd.)

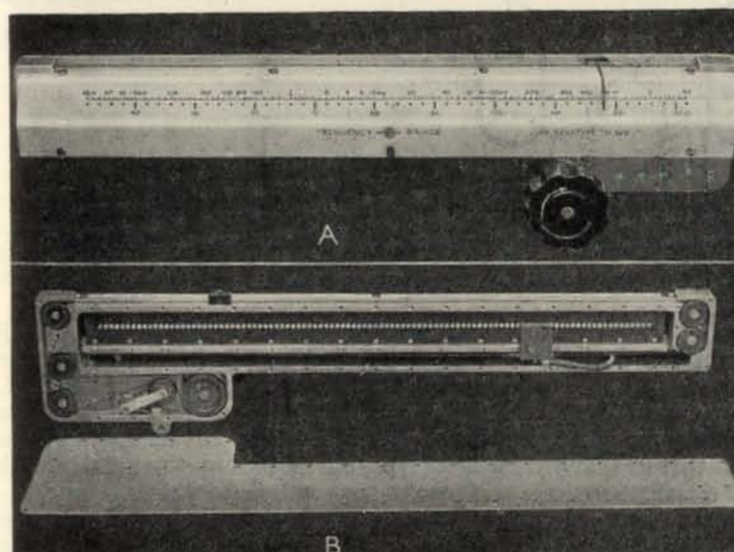


Fig. 4.31—Front (a) and back (b) of 100-db attenuator used in TF.867 standard-signal generator. An additional 40 db is obtainable with the terminating unit seen at the foot of Fig. 4.30. (Marconi Instruments Ltd.)

4.31. 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 and pulse, shown as *a* and *b* respectively in Fig. 4.32. 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.

In one main class of methods a more or less sinusoidal wave is squared by passing it into a limiting amplifier so that the sides are steepened and the tops sharply cut off. One such device is a valve with a very short grid base (to cut off the negative half by bottom bend) and a high resistance in series with the grid (to cut off the positive half by grid current). The resulting approximation to a square wave can be converted into a pulse by means of a differentiating circuit—a grid leak and capacitor having a time constant much less than the period of one cycle. The resulting peaked signal can, if necessary, be squared

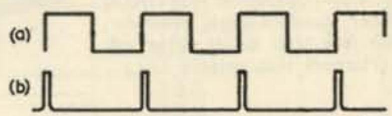


Fig. 4.32—Square-wave (a) and pulse (b) waveforms

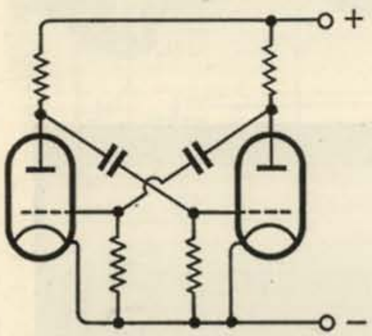


Fig. 4.33—Basic multivibrator circuit

off by further limiters. This technique is simple and straightforward, but a large amount of amplification is needed to make the wavefronts really steep.

The alternative is to generate a steep-fronted signal directly, usually in some RC version of one of the types of oscillator already described in this chapter. Some of these are self-repeating (but can be " locked " to an alternating signal within certain limits of frequency); others function only when triggered by a pulse. Some of these circuits are rather trickier than the distorter type, but a closer approximation to the ideal rectangular shape can be obtained with fewer valves.

Details of such circuits are given in any book on radar. 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

* The sawtooth waveform used for time bases is considered in connection with oscilloscopes in Sec. 5.27.

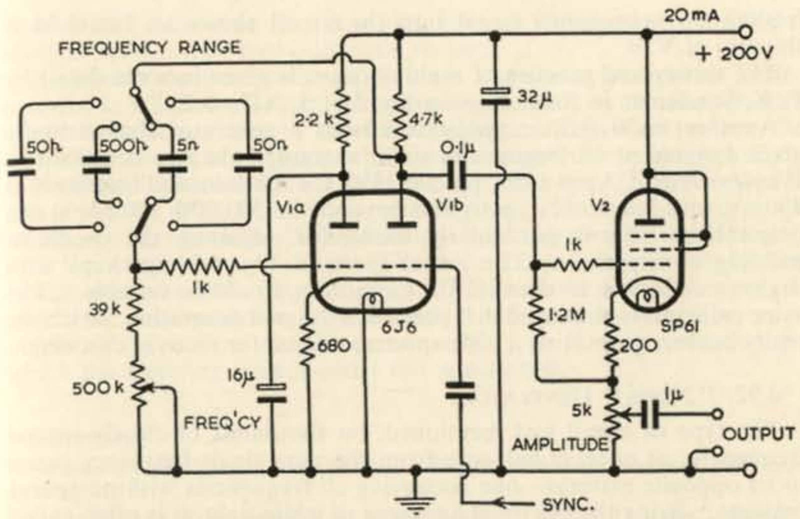


Fig. 4.34—Circuit diagram of 15 c/s to 160 kc/s square-wave generator

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 in *Wireless World*, January 1951, p. 35. 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. 4.33. 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. 4.15) and is derivable from Fig. 4.16 by shorting out R_1 and omitting C_2. A simple and inexpensive square-wave generator based on this circuit, variable from 15 c/s to 160 kc/s, is described by L. F. Sinfield in *Wireless World*, July 1952, pp. 285–6. The full circuit is shown here as Fig. 4.34. 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.

A two-valve circuit for developing 1-μsec pulses of 30 V amplitude and <0·5 μsec rise time is given by F. A. Benson and G. V. G. Lusher in *Wireless Engineer*, January 1952, pp. 12–14.

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. 4.34 is the grid of V1b.

The theory and practice of multivibrators is given in some detail by E. K. Sandeman in *Radio Engineering*, Vol. 1, XII: 6.2.

Another multivibrator application is as a generator that virtually gives a signal at all frequencies simultaneously. In one described in *Wireless World*, April 1939, pp. 349–350, 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. 4.34, 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.

4.32. " 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 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

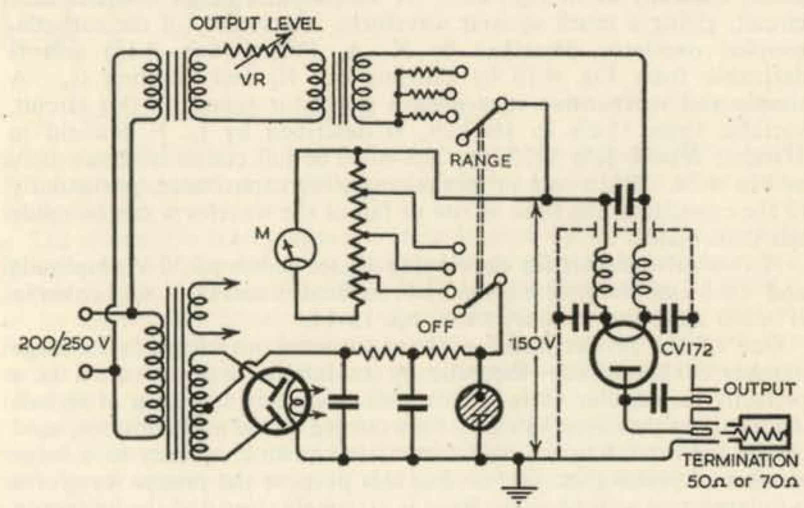


Fig. 4.35—Circuit diagram of Marconi Instruments noise-voltage generator TF.987

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 make the noise level much more difficult to calculate. The Marconi Instruments TF.987 embodies this principle; Fig. 4.35 is its circuit diagram. The noise-factor range switch controls both the filament voltage (and therefore its temperature) and the anode current shunt. The noise output is adjusted by the continuously-variable filament control VR, 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 50-Ω or 70-Ω termination to which the receiving system under test is matched.

References:

" Noise Factor ", by L. A. Moxon. *Wireless World*, December 1946 and January 1947.

" Theory and Measurement of Noise Factor ", by R. J. Yates. *Marconi Instrumentation*, July–August and October–November 1950.

" A Generator of Electrical Noise ", by A. P. G. Peterson. *General Radio Experimenter*, December 1951, pp. 1–9.

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 valves; (2) valve 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

Normal current meters are 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.

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).

The 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.*, September 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 per cent second harmonic would be about 1·1 per cent, but the same percentage third harmonic would cause anything up to 5 per cent error. In *Wireless World*, August 1951, pp. 316–9, Thomas Roddam shows how to combine rectifiers with resistors to make a nearly square-law voltmeter, in which waveform error is practically eliminated; and he describes two methods of testing the accuracy of square law.

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.

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 unsuitable, 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. Multi-range Meters

There is now a very wide choice of multi-range meters, reading direct current and voltage. Most of them also have alternating-voltage ranges, many have alternating-current ranges, and often they have some resistance, capacitance, etc., ranges. An example is shown in Fig. 5.1. The nucleus is commonly a moving-coil meter fully deflected by 1 mA, with higher current ranges obtained by shunts, and voltage ranges by series resistances (multipliers) of 1,000 Ω per volt. To enable interchangeable external shunts to be used for heavy currents, there is a standard full-scale voltage drop: this is 0·075 V, unless otherwise marked. Since 1 mA is too high a meter current for accurate voltage readings in valve circuits, an alternative standard type is a 50 μA meter (20,000 Ω per volt). In theory, the construction of shunts and multipliers for providing desired ranges is simple; in practice it demands rather more care for a given standard of accuracy than the uninitiated might suppose. This work is dealt with by

C. R. Cosens in *Wireless Engineer*, November 1934.

For a.v. ranges above about 25–50 full scale, the same scale as for d.v. can be used, provided that (for the reason explained in Sec. 5.2) the multiplier resistances are 10 per cent less than for d.v., or (more conveniently) the sensitivity of the meter is increased 10 per cent by unshunting. Owing to the varying resistance of the rectifier, the method of obtaining current ranges by shunts cannot be used for a.c., and it is necessary to have a current transformer. Its construction is not difficult, but to anyone accustomed to ordinary voltage transformers its design is very unorthodox. A comprehensive booklet (MR3) on the use of meter rectifiers is obtainable from the Westinghouse Brake and Signal Co., Ltd. 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).

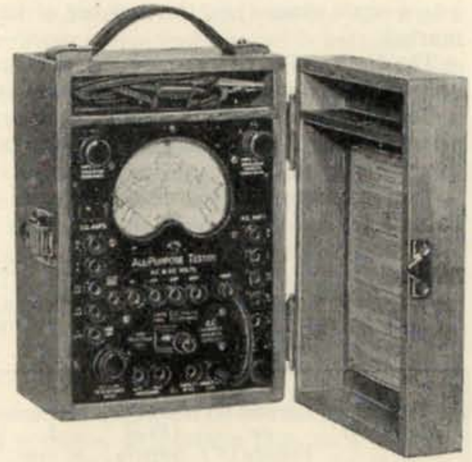


Fig. 5.1—"Radiolab All-Purpose Tester", reading in volts, milliamps, ohms and microfarads over wide ranges. (Everett, Edgcumbe & Co., Ltd.)

5.4. 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, laid down in B.S. 89,* which is good enough for most ordinary work. Because these descriptions are rather misleading they have been altered in the new edition of B.S. 89 (1954) 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 $\frac{1}{3}$ for electrostatic and thermal types, except that in electrostatic voltmeters having smaller than $3\frac{1}{4}$-in scales it is from

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

¼ to ¾ 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 per cent of the scale range for moving-coil and 0·5 per cent for all other types. For self-contained multi-range ammeters it is 0·5 per cent for all types.

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

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 Switchboard (S)*	*Scale length in inches*			
			5 or over	3¼ up to but not including 5	2 up to but not including 3¼	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	—

There is an extra allowance of 0·25 per cent for multitrange instruments, and 1 per cent 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 per cent for Precision meters and 0·10 to 0·30 per cent 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 1·5 + 0·25 per cent 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 0·15 × 20 = 3 volts. So even an up-to-standard moving-coil instrument with a reasonably large scale can be very 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 is common) may not often have to be allowed for, but the current taken during voltage readings should be (Sec. 10.1). This 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 are 1, 2 and ½, combined if need be with multiples or submultiples of 10.

5.5. The Rectifier Automatic Shunt

Milliammeters used for general testing, even in the hands of careful workers, run the risk of bent pointers, or a worse fate, as a result of accidental excess currents, unless fitted with a reliable cut-out. By the time a low-resistance shunt is switched across, the damage is done. What one wants is an automatic shunt that adjusts itself according to the current. A metal rectifier is such a shunt. Things can be arranged so that when the current is small its resistance is very much higher than that of the milliammeter across which it is connected, thus preserving practically the full sensitivity of the meter near the zero end of its scale. But as the current increases, the rectifier shunt resistance drops; with a suitable choice of rectifier an approximately logarithmic scale can be obtained over a very wide range—as high as 1 A, using a 5-mA meter. Unfortunately, individual rectifiers even of the same type are liable to differ considerably in this shunting characteristic, which also changes somewhat with temperature and with age, so the rectifier-shunted meter is not an instrument of precision but is useful for obtaining approximate readings of currents that are liable to vary widely, or for signal indicators, or for null-reading indicators such as in bridges.

Normally the resistance of the meter has to be increased in order to bring the rectifier on to a suitable range of its characteristic, and to avoid needless voltage drop no rectifier disks should be in series. Fig. 5.2 shows how to connect a Westinghouse 2-1-4.BLC rectifier to provide four rectifier paths in parallel in the active direction and four in the reverse direction to protect against reverse currents. Fig. 5.3 shows the very nearly logarithmic scale up to 250 mA obtained with a sample of this type of rectifier. The series resistance is adjusted to give the desired full-scale reading, and in this example brought the total meter resistance up to about 70 Ω and the full-scale voltage drop to 0·35 V.

For protecting sensitive meters the characteristics of germanium rectifiers are particularly suitable. Selenium and copper oxide rectifiers can be used, however, as specified in an informative article

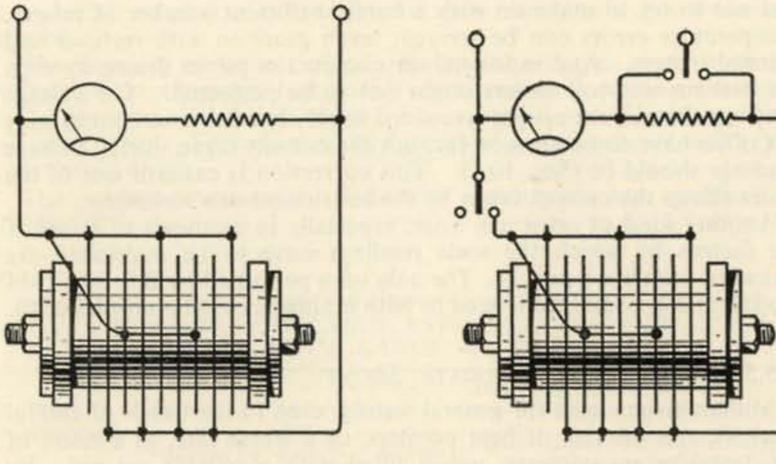


Fig. 5.2—(a) Method of connecting Westinghouse 2-1-4.BLC rectifier as a shunt. The series resistance is adjusted to give the desired full-scale reading; the scale shape then depends on the rectifier characteristics. (b) Addition of switch arms for reverting to unshunted meter when desired

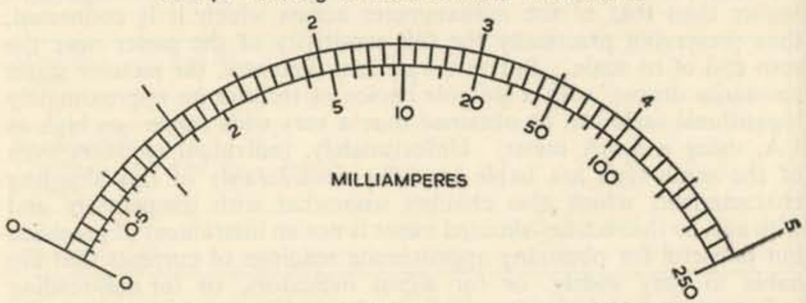


Fig. 5.3—Scale of rectifier-shunted milliammeter compared with original to show the great increase in range of current measurable, and the logarithmic scale shape

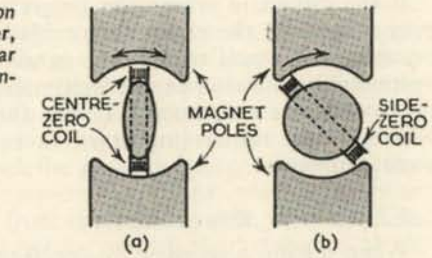
on the protection of microammeters, by J. de Gruchy (*Wireless World*, September 1953, pp. 425–7).

An alternative method of protecting the pointer of a moving-coil galvanometer is to use a pole-piece shaped somewhat 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 to give a uniform strength at all coil positions and hence a linear scale. This method does not, of course, protect the coil from being burnt out.

The risk to a sensitive instrument is particularly acute in high-voltage testing, and a method suggested by T. E. Burnup,* effective up to thousands of volts, is shown in Fig. 5.5, which explains itself.

* *Electronic Engineering*, April 1952, p. 184.

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



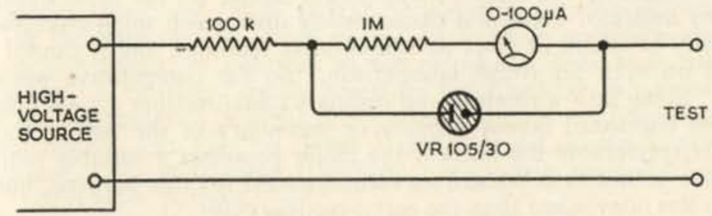


Fig. 5.5—Current-limiting device for protecting microammeter used to test current leakage at high voltages

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 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 radio 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 in connection with the total power taken by a receiver there are other ways (Secs. 10.5 to 10.6) of measuring power factor.

The foregoing are the fundamental pointer instruments; there are now many 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, it is possible to get some idea of the actual power in it 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 valves that the load resistance is adjusted at least approximately to the optimum value, or whatever other value may be required for

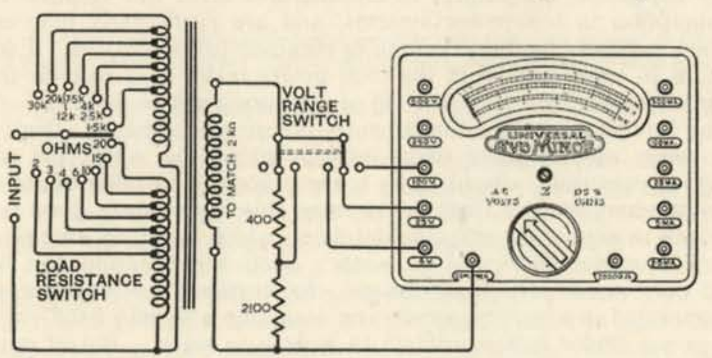


Fig. 5.6—Circuit of unit for adding to multi-range test set to convert it into an output meter

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. The difficulty is avoided by means of a transformer with a fixed secondary, across which the voltmeter and a suitable fixed load resistance are connected; and the primary is tapped, so that the fixed load resistance can be made equivalent to any one of a number of different resistances, looked at from the primary side. Under these conditions a transformer to give nearly perfect characteristics at all frequencies from, say, 30 to 10,000 c/s is a difficult piece of design; but for the purposes for which an output meter is generally used a very high standard of accuracy at the extreme conditions is hardly necessary. A relatively simple accessory for attaching to a rectifier-type a.c. voltmeter is a great improvement on the single-resistance output meter; and Fig. 5.6 shows a circuit based on the Universal Avominor. The multi-range transformer is specially made by Sound Sales Ltd. for the purpose.

The selection of seven low and seven high resistance ratios is enough to enable output characteristics to be plotted. The average loss due to the transformer is 1 db, increasing somewhat above 6 kc/s. The secondary winding is designed for the resistance of the Avominor at its lowest range—2 kΩ. On the other ranges the resistance is higher, so extra shunt resistors are needed to maintain the load constant. The meter reads as low as 0·1 milliwatt; and, since 5 W is about as high as is needed in most work, only the first three ranges of the Avominor are utilized. It should be noted that on the 5-V range the resistance of the meter rises much above 2 kΩ at small deflections, so tests in which this matters should be run at not less than 5 mW. The transformer, resistors, and range switches can be built into any compact form that suits the experimenter. The resistors should be non-inductive; the composition type is good enough if selected and checked for correct resistance within a few per cent. The 2,100-Ω resistor has to dissipate up to 5 W, so it may be necessary to use

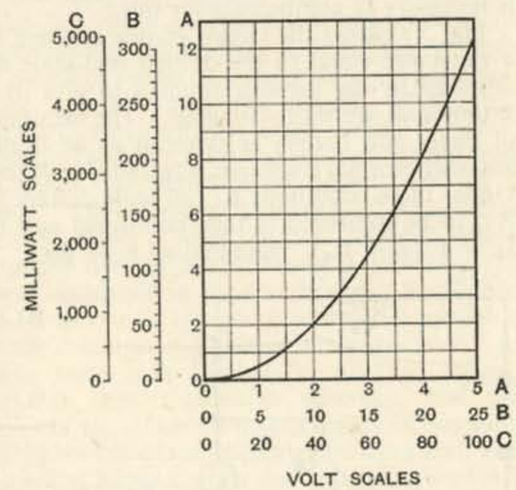


Fig. 5.7—Curve for converting volt scales to milliwat scales, based on a load resistance of 2 kΩ

several resistors of lower rating connected in series or parallel. An alternative is to use the wire-on-mica type described in Sec. 6.4. To make the instrument read power directly without interfering with the existing scale one can cut a piece of stiff Bristol board into a shape to fit snugly over the glass without sliding about. A curved slot is cut to reveal the position of the pointer over the whole of its movement, and the milliwatt scales marked above and below. Fig. 5.7 is useful for deriving the milliwatt scales from the volt scales.

In the Dawe 610B output meter a compensating system is used to equalize readings over the wide range of load resistances included, from $2\frac{1}{2}$ to 20,000 Ω. To overcome the difficulties associated with the transformer, a purely resistive Π attenuator is used as the impedance-matching device in the Airmec Type 708 output meter.

A variable-impedance 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. Many 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—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. 5.8 shows the circuit of one of these simple ohmmeters. R is a resistance equal to the desired mid-scale reading, and the meter is 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

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

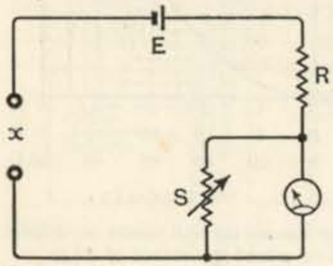


Fig. 5.8—Circuit of simple ohmmeter. The resistance to be measured is connected to the "x" terminals

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 are therefore immaterial it is allowable to use the variable shunt, as described, for bringing the pointer initially to full deflection (zero on 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. 5.9 shows $I_1/I_2 - 1$ plotted along a linear 0–I_1 scale, and is a universal resistance scale for this type of instrument, the reading only

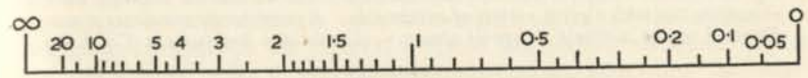


Fig. 5.9—Scale of the Fig. 5.8 type of ohmmeter, requiring only to be multiplied by R to indicate the resistance connected at "x"

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 in *Wireless World*, August and September 1943; Cazaly and Roddam in *A.C./D.C. Test Meters* (Pitman); and W. Tusting in *Wireless World*, July 1953.

Even with more than one scale it is difficult with one instrument to cover anything like the full range of resistances encountered in radio 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. The most satisfactory method is to use the unknown resistance as a shunt to a current meter, the reduction in deflection being a measure of the resistance. The Ferranti low-reading ohmmeters depend on this principle. 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. 5.10, 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 0·001–1,000 Ω with reasonable accuracy, and the scale factors are all multiples of 10. There is no possibility of bumping 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.

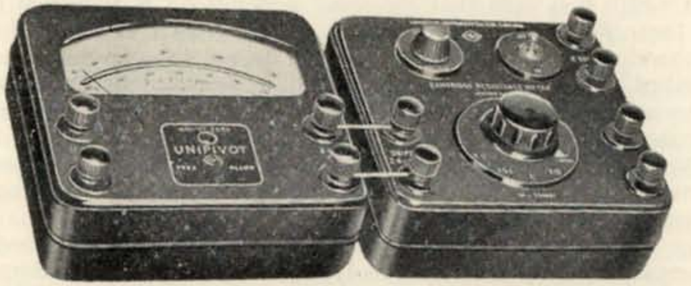


Fig. 5.10—The " Unipivot Versatile Galvanometer " can be used for accurate work in conjunction with a great variety of accessories. A multi-range ohmmeter manufactured to the author's design is shown. (Cambridge Instrument Co. Ltd.)

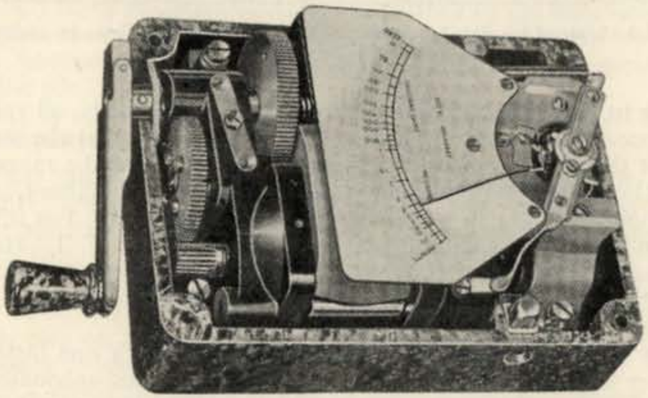


Fig. 5.11—" Megger Tester Series 3", a direct-reading ohmmeter for insulation resistance, incorporating a 500-V hand-driven generator. The instrument is shown with cover removed. (Evershed & Vignoles Ltd.)

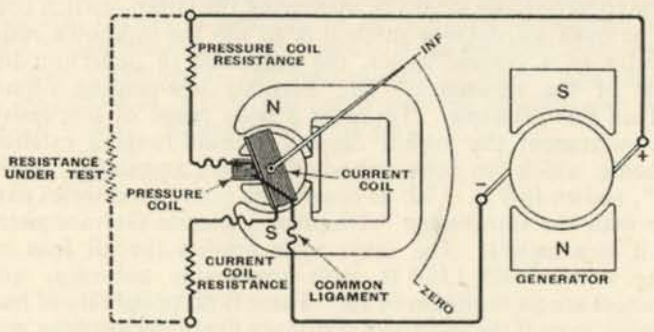


Fig. 5.12—Circuit of the " Megger Tester Series 3". With infinite resistance, current flows only through the pressure coil, causing the pointer to be held in the full-scale position. Any current flowing through the resistance under test traverses the current coil and causes a corresponding deflection. (Evershed & Vignoles Ltd.)

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 rather large 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. 5.11 and 12 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 0·25 μF. Testing of large capacitors is impracticable with this model owing to the charge and discharge currents as the generator speed fluctuates, but other models are obtainable with a special constant-output generator for high-capacitance circuit tests, and 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. 5.15.

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 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. 5.20).

5.10. Valve Voltmeters: Desirable Characteristics

It is sometimes assumed that valve voltmeters are necessarily alternating-voltage meters, and might in principle be classed with the other rectifier types already described. But the most fundamental characteristic of valve voltmeters is not rectification; it is what might be called impedance transformation. In radio and electronic work there are very many applications where the ordinary meter would draw too much power or otherwise unduly disturb the circuit under test if it were connected directly to it, and what is needed is something between the circuit and the meter, not appreciably affecting the circuit, yet giving out the right amount of power to deflect the meter to the correct reading. It is the fact that valves can readily be arranged to have a very high input impedance and relatively low output impedance that makes them so valuable for this purpose. Clearly the ideal would be infinite input impedance (for then there would be no effect at all on the voltage being measured) and zero output impedance (for then the calibration of the meter would not be affected). An important supplementary clause is that readings should not be appreciably affected by ordinary variations in the valve or valves and their supply voltages. A design approximating closely to this ideal is described in Sec. 5.14. Other desirable features are wide range of measurement, low cost, portability, ease of use, and unbreakableness.

These requirements for d.v. voltmeters hold also for a.v., but virtually infinite input impedance is much more difficult—one might say impossible—to achieve for a.v., especially as another requirement is usually a very wide range of frequency. Also, a.v. measurement is greatly complicated by the question of waveform and " values ". For most purposes one would like to know r.m.s. values. Unfortunately this is much more difficult and inconvenient to arrange than half-cycle average values, which are seldom required. Sometimes peak values are required, and this is fairly easy to arrange, though not very accurately with extreme waveforms. Instruments that respond to one value and are calibrated in another are correct for only one waveform and can be very misleading. With the extensive use of non-sinusoidal waveforms, the tendency has been to displace valve voltmeters by cathode-ray equipment, so that the waveforms can be seen as well as measured; but for many purposes a valve voltmeter is more convenient. No one kind can fulfil all requirements, however, so it will be necessary to consider a number of different types.

Because a lot of rather exacting requirements have just been mentioned it should not be concluded that any valve voltmeter worth considering has to fulfil them all closely. To secure an accurate and permanent calibration, much thought and trouble must be expended on the design. But again, the fundamental thing is the impedance transformation, and if this is obtained it may be possible so to arrange the measurement that only very limited calibration, or even none at all, will do.

For instance, the detector valve in any receiver can be made into a valve voltmeter by inserting a meter in its output circuit, and although this is a very cheap and simple device it fulfils the most important requirement by not disturbing the signal circuit at all. The need for calibration is avoided altogether if the valve voltmeter is used only to indicate equality of voltage. The next best thing is to make the result of the experiment depend on the two voltages being not too widely different, so that error tends to cancel out. An example of this is a certain method of measuring r.f.

resistance (Sec. 9.26), in which the duty of the valve voltmeter is to indicate a voltage ratio of $1 : \sqrt{2}$. So although a valve voltmeter with an accurate and stable voltage calibration is very desirable, it is possible to overestimate this requirement.

5.11. Survey of Valve Voltmeter Types

Before going into details, it may be helpful to survey the established types of valve voltmeter. Like the addresses in telephone directories, the circuit diagrams in Fig. 5.13 are for identification only and barely include the essentials. Nearly all a.v. valve voltmeters are elaborations or variations of these six main types. V denotes the alternating voltage being measured.

(*a*) *Diode*. C charges up to the peak value of V, and this is measured by the meter, which ought to be a sensitive microammeter and include a resistance R many times greater than that of the diode, otherwise the voltage of C is substantially lowered and a large current taken from the source of V. Logically this type should be included with the ordinary rectifier meters in Sec. 5.2, because the use of a thermionic rectifier in place of one of the copper-oxide or germanium type introduces no new principle, and the power to operate the meter comes entirely from the source of V, so the essential feature of a valve voltmeter (as here considered) is lacking; but *a* is shown because it is the seed from which *b* and *c* developed.

(*b*) *Cumulative Grid*. Here the grid and cathode of the valve rectify V as in *a*, and the resulting direct voltage is amplified by the valve as a whole. Since the valve draws no grid current except at the peak of the positive half-cycle, and that can be minimized by making R very large, there is a considerable degree of impedance transformation, and the meter can be an ordinary milliammeter while loading V only slightly. This was one of the first types of valve

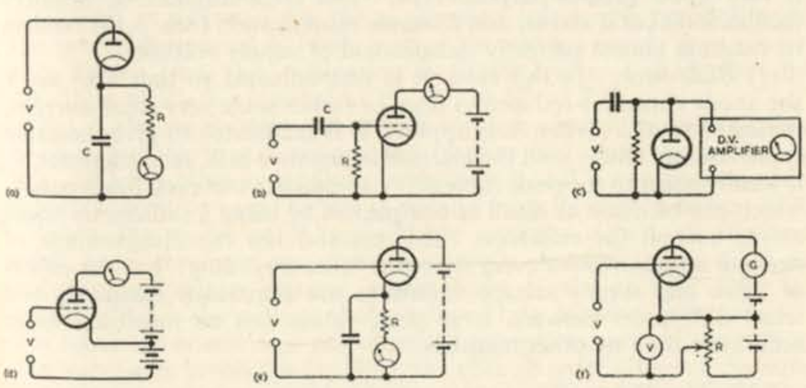


Fig. 5.13—Six basic types of valve voltmeter: (a) *simple diode rectifier;* (b) *cumulative grid, equivalent to diode plus amplifier;* (c) *diode with separate amplifier;* (d) *anode-bend;* (e) *reflex;* (f) *slide-back*

voltmeter, and is one of the most sensitive, but has largely gone out of favour because the meter reading is greatly affected by variations in the valve and its anode and heater voltages. Moreover the range of voltage measurable is very limited, and the scale corresponds to no simple law. The current through the meter is maximum when reading zero, and decreases with increasing V, so the instrument is protected against overloading. In some elaborations, however, provision is made for balancing out the " zero " current.

(*c*) *Amplified Diode.* To avoid the disadvantages of *b*, a separate amplifier is used for the d.v. output of a diode, instead of making one valve serve as both. The majority of modern valve voltmeters are of this type, the d.v. amplifier usually being a balanced system with negative feedback.

(*d*) *Anode Bend.* All the foregoing types have to draw enough current from the source of V at its positive peak to keep current flowing through R during the whole cycle, so the input resistance is certainly not infinite. The remaining three types are attempts to overcome this. In *d*, the valve is biased to the bottom bend, so the application of V causes the anode current to increase. This type requires stability of grid bias as well as anode voltage and it is much less sensitive than *b*. Within its equally limited range of V its input resistance is much higher, and it has one important claim to consideration for some purposes: with a suitable choice of valve and operating conditions it approximates closely to a square law, so reads true r.m.s. values, and the waveform errors are much less than in other types.

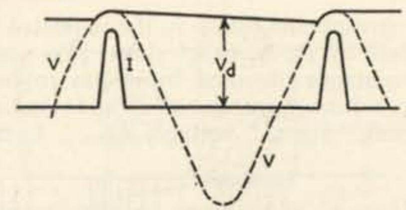
(*e*) *Reflex, Cathode-follower, or Auto-bias.* Here the negative grid bias is not fixed, as in *d*, but increases with increasing V; consequently a much greater range of V is measurable, and the scale is very nearly linear. The reading is proportional neither to peak nor mean value, but something between the two. Where this does not matter, this is a very good general-purpose type. The large amount of negative feedback makes it stable, and a simple modification (Sec. 5.19) renders its readings almost perfectly independent of supply voltage.

(*f*) *Slide-back.* In this type, R is first adjusted so that with no V the anode current is reduced to zero, or rather some very small current, indicated by G. When V is applied, R is readjusted to give the same small reading on G; and the increase in negative bias, read on meter V, is ideally equal to the peak value of V. In practice there is a discrepancy, which can be made as small as one pleases by using a sufficiently small anode current for reference. This method has the disadvantage of needing an adjustment every time one takes a reading; but the effects of valve and supply-voltage variations are practically excluded, and small differences between large peak values can be measured more accurately than by other methods.

5.12. The Diode Rectifier

The diode rectifier of Fig. 5.13*a* is, in one form or other, the essential feature of types *b* and *c* also. Because it looks so extremely simple

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



it might be supposed that it calls for little consideration. But rectifiers are much less simple than they look, and even after several decades of use the diode rectifier is often misunderstood. The following is an outline of the author's article in *Wireless World*, March 1952, pp. 89–94, which was an attempt to clarify the workings of the diode in so far as they concern valve voltmeters. See also June and July issues, 1954.

If the diode were a perfect rectifier and R were omitted, 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 per cent of each cycle and only $10/2 = 5$ kΩ during the remaining 5 per cent. 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*a* necessitates a d.c. path through the source, and as this is often lacking in a.c. circuits the rectifier is usually rearranged as at *c*. 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. 4.10) 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. 5.15, adapted from data by D. A. Bell,* 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-tenth of R it

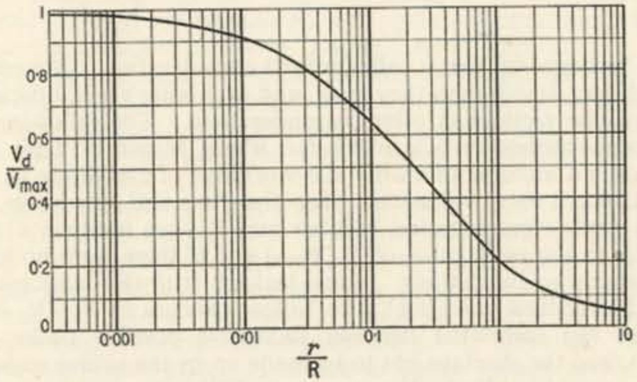


Fig. 5.15—Effect of the ratio r/R (input and diode resistance to load resistance) on efficiency of diode rectification

causes an error of nearly 10 per cent, and if it were equal to R the voltmeter would cut the voltage down to one-fifth!

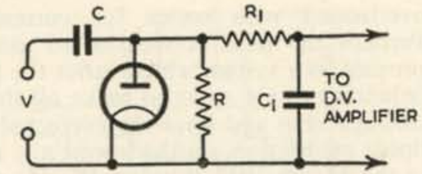
The effects of this on valve-voltmeter design are far-reaching. Obviously in Fig. 5.13*a* R cannot be made enormously large if there is to be a reasonable meter current. The object of the important *c* type is to enable a very high R to be used; but unless special precautions are taken, as explained in the next section, it may not be possible to raise it even to 1 MΩ without causing appreciable error, due to amplifier input conductance.

Assuming 1 MΩ, however, Fig. 5.15 shows that an *r* of 4 kΩ causes 5 per cent error, which on a 100-V range is greater than that due to a 1,000 Ω/V straight voltmeter! So, for work on non-resonant circuits 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.

Even the forward resistance of the rectifier 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.

* " Diode as Rectifier and Frequency-changer ". *Wireless Engineer*, October 1941, pp. 395–404.

Fig. 5.16—The influence of the values of the components marked here on the performance of the diode rectifier is discussed in the text



It should be noted that the change from *a* to *c* in Fig. 5.13 causes V_d to be accompanied by the full alternating voltage V as well; and to prevent this from overloading the 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. 5.16, 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 transferred to C_1, but V_d is stepped down. (A $\sqrt{2}$: 1 step-down 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 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 cannot 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 very skilful design to avoid drastic reduction in accuracy or frequency range or both. The Philips GM.6006 voltmeter is an example of what can be done; it is essentially an amplifier with a stable voltage gain of × 500 over the frequency band 1–30,000 kc/s, and the full-scale reading on its lowest range is 1 mV. Full details are given in *Philips Technical Review*, January 1950, pp. 206–214; see also end of Sec. 6.22. But even with the most modern technique one could hardly hope to cover 20 c/s to 200 Mc/s, which is a normal range for a diode voltmeter. If the frequency range is limited the difficulties of pre-amplification are correspondingly reduced; S. Kelly gives constructional details of an a.f. valve voltmeter reading 1 mV to 10 V, in *Wireless World*, June 1951, pp. 215–218.

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 rather 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 *Wireless World* article mentioned, and an example is given in Sec. 5.14.

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. 5.21). To keep the error due to it below about 1 per cent, the time constant CR should be at least 20 times the period of the lowest frequency; say 1 (megohm-microfarads) for 20 c/s.

Supplementary references:

"The Diode as A.C. Voltmeter", by C. S. Bull. *J. Sci. Inst.*, October 1947, p. 254.

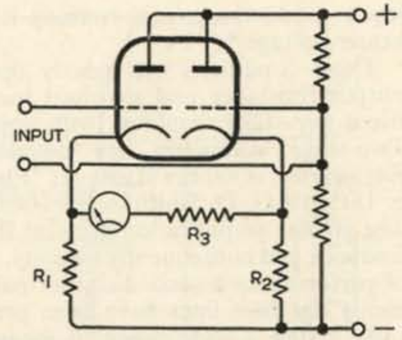
"Diode Valve Voltmeter Errors", by G. D. Morgan. *Electronic Engineering*, December 1952, pp. 575–7.

5.13. The D.V. Amplifier

An advantage of type *c* (Fig. 5.13) is that the amplifier and meter on their own can be used for d.v., and plugging in the diode rectifier adapts it for a.v. If, as suggested in the last section, the rectifier filter reduces the rectified voltage in the ratio of approximately $\sqrt{2} : 1$, the d.v. scale reads also r.m.s. a.v. (of sine waveform), subject to suitable arrangements being made for the bottom bend.

The disadvantages of a simple single-valve d.v. amplifier are that when working reasonably linearly there is a large initial anode current to flow through the meter, and that this current varies largely with valve and supply 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 shown in outline in Fig. 5.17, is the basis of most present-day commercial valve voltmeters. Constructional details of an instrument of this kind are given by S. W. Amos in *Wireless World*, December 1950, pp. 430–2. An advanced example is the General Radio Type 1800A, described in *General Radio Experimenter*, September 1946. Alternatively the load resistors and meter can be connected on the anode side. The relative merits of these alternatives, and of the corresponding single-

Fig. 5.17—The cathode-loaded form of the twin-triode d.v. amplifier commonly used in valve voltmeters



valve voltmeters, are discussed by P. Popper and G. White in *Wireless Engineer*, December 1948, pp. 377–384. See also J. D. Clare in *Wireless Engineer*, July 1948, on the type shown in Fig. 5.17. The second valve does nothing to increase the sensitivity, reckoned in meter-current change per volt change; in fact, it halves it. But by automatically balancing out a large part of the undesired variation it enables a more sensitive meter to be used, thereby increasing sensitivity reckoned in terms of deflection. The advantage of cathode connection of the meter 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 to be large in comparison with this. The resistance in series with the meter therefore consists of its range-adjusting multiplier, R_3, and twice the valve output resistance. 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 proportion of the meter-circuit resistance will be in the form of "electronic" resistance rather than the more reliable wire-wound sort.

Ideally, the amplifier should have zero output resistance, so that the meter can work entirely with stable multiplier resistances. Even with the highest g_m valves their resistance in the Fig. 5.17 types of circuit can hardly be reduced below hundreds of ohms. And it is precisely this kind of valve that is least likely to fulfil the other half of the ideal—zero input conductance. It is not always realized that merely applying negative bias is not good enough for accurate measurement purposes. Comparison of anode current with and without a high resistance—say 20 MΩ—between grid and bias demonstrates this very clearly. As much as 1·5–2 V negative bias is needed to cut off ordinary grid current; beyond that there is grid 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 sufficient to make the grid 0·2 V more positive when 20 MΩ is in the grid lead!

There are special electrometer valves designed for extremely high input resistance. Without going to the expense of these, one can get results several orders better than usual by choosing a type such as the EF37A or EF86, which is known to have exceptionally low input conductance, and running it with low anode voltage and low heater voltage.*

These conditions are exactly opposite to those required for low output resistance and a robust inexpensive meter. The answer is to use a two-stage amplifier, with appropriate input and output valves. Two-stage voltmeters are described by R. Kitai in *Electronic Engineering*, October 1950, L. Fleming in *Electronics*, April 1951, p. 181, and G. D. Smith in *Electronic Engineering*, January 1952, p. 17. The greater amplification presents the opportunity for greater negative feedback and consequently stability. To maintain the highest standard of performance a second similar pair is needed as a balance. Instruments on these lines have been produced by Electronic Instruments Ltd., giving a wide range of measurement and accuracy previously associated only with high-grade valveless meters. Another, due to C. Morton, is specified in British Patent 636,212.

5.14. A High-stability Valve Voltmeter

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

(1) Input conductance less than 10^{-10} mho (with under-run EF37A 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 0–1·5 V range using a 0–3 mA movement;

(3) Output voltage equal to input voltage within a fraction of 1 per cent, 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.

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

A full account of this type appears in *Wireless World*, January 1952, pp. 14–18, and a briefer one in *Electronics*, December 1951, p. 142. Fig. 5.18 is the basic circuit. Each half of the amplifier is in effect a voltage stabilizer of the Fig. 4.2 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. 5.19 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

* "H.F. Pentodes in Electrometer Circuits", by K. D. E. Crawford. *Electronic Engineering*, July 1948, pp. 227–231.
"Receiving Valves Suitable for Electrometer Use", by G. A. Hay. *Electronic Engineering*, July 1951, pp. 258–261.

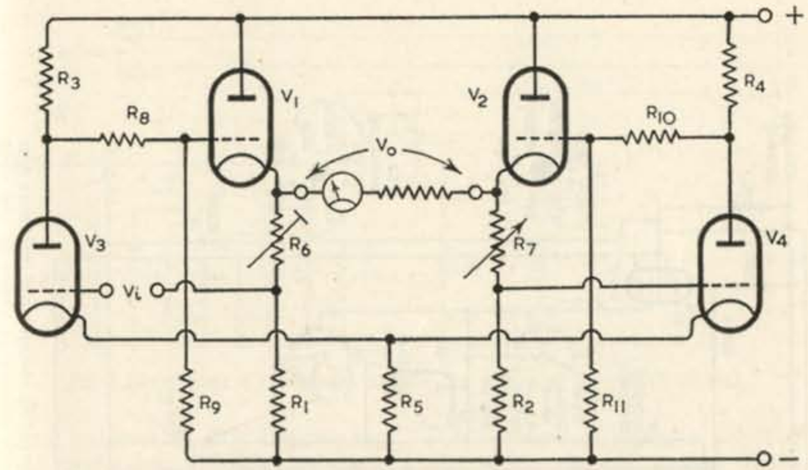


Fig. 5.18—Theoretical diagram of circuit for providing any ordinary voltmeter with virtually infinite input resistance without affecting calibration

practically perfect linearity. Using a 0–3 mA meter, ranges with full-scale readings of ±1·5, 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., described in Sec. 5.12. 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.)

5.15. Valve Ohmmeters

A valuable feature of a d.v. valve voltmeter is the ease with which wide-range ohmmeter facilities can be incorporated. Fig. 5.19 shows provision for mid-scale resistance ranges of 1, 10, 100 and 1,000 kΩ, applicable to the scale of Fig. 5.9, giving satisfactory reading from 100 Ω to 10 MΩ and rough readings over an additional decade at 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.

An example of a versatile commercial instrument providing resistance and insulation ranges as well as d.c., d.v., a.v., power, and capacitance, is the Avo electronic test meter.

If very high resistances are to be measured it is generally better to design a special instrument for the purpose. Obviously the choice of

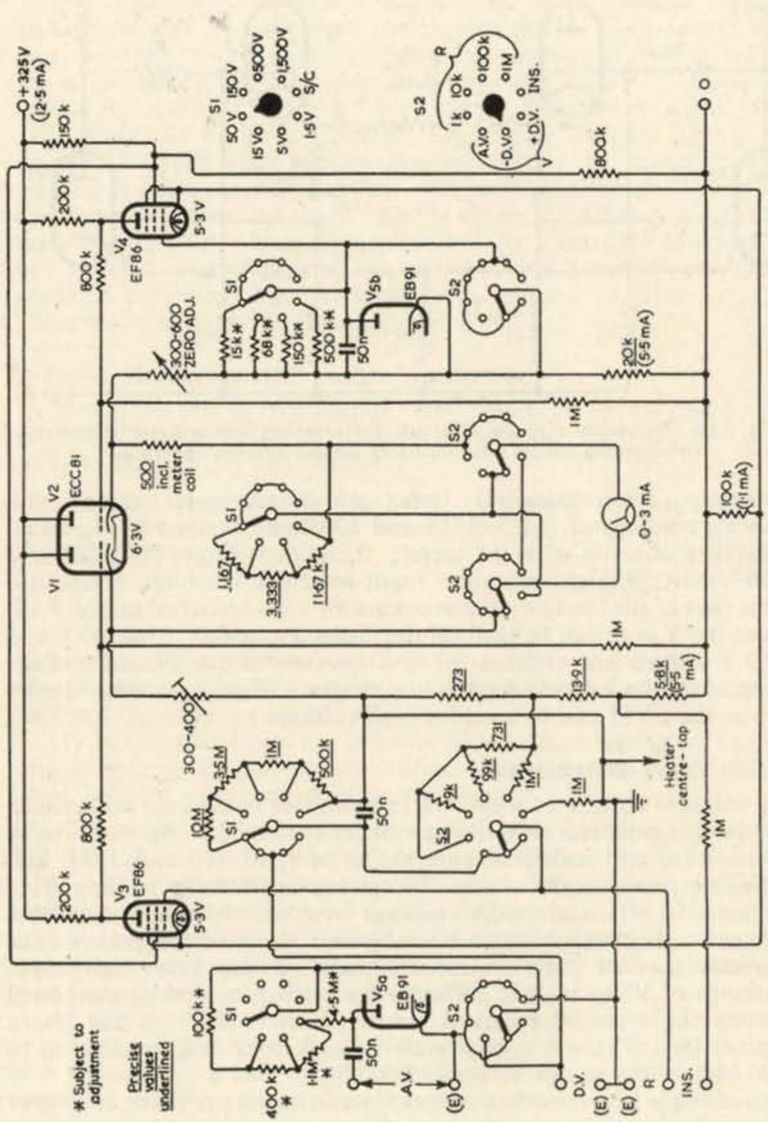


Fig. 5.19—Practical development of Fig. 5.18 for measuring direct and alternating voltages and resistance over a wide range. S1 and S2 each comprises a single ganged switch, with settings as shown on the right

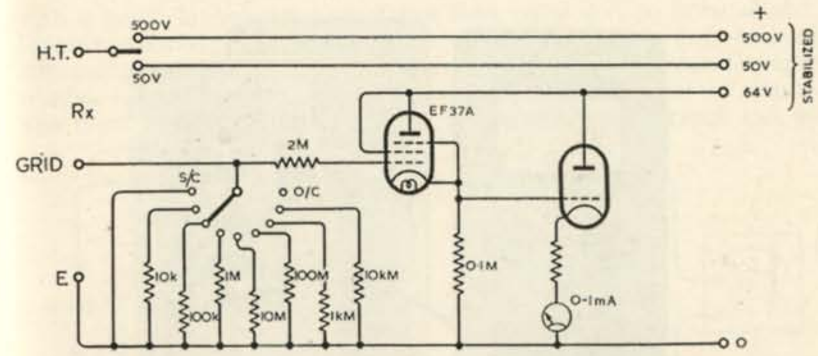


Fig. 5.20—Circuit of valve megohmmeter for measuring 0·1–1,000,000 MΩ

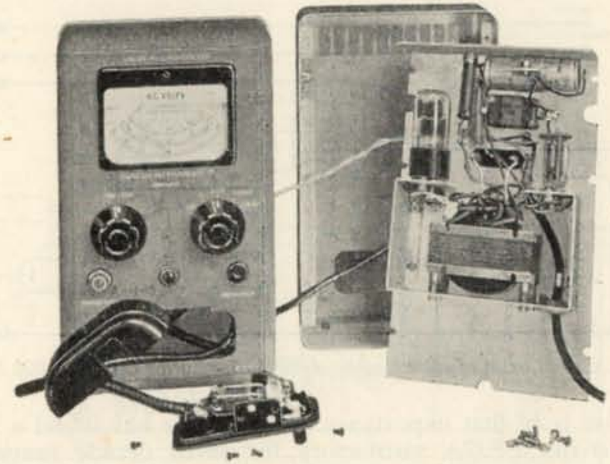
input valve is of first importance. The author has found a selected and under-run EF37A satisfactory for seven decade ranges from 0·1 MΩ to at least 1 MMΩ with rough indications up to 10 MMΩ (10^{13} Ω). The instrument is, in effect, a 0–5 V d.v. voltmeter; its main features are shown in Fig. 5.20*. 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.

The design and use of electronic megohmmeters are discussed by H. G. M. Spratt in *Wireless World*, October 1948, and details of a General Radio instrument reading 0·5 MΩ to 2 MMΩ are given in *General Radio Experimenter*, November 1951. For a simple low-reading megohmmeter (5 kΩ–5 MΩ), see W. H. Cazaly in *Wireless World*, September 1949, pp. 326–8.

5.16. The Cumulative-grid Valve Voltmeter

This type, Fig. 5.13*b*, is just a leaky-grid detector, and has largely gone out of use because of its poor stability, large initial current, and limited range; also its high input loss compared with types *d* to *f*. But this is perhaps unfair, for it has the advantages of simplicity and high sensitivity, and the balancing methods used for stabilizing type *c* are available for mitigating its disadvantages. And the meter is safe, for the " zero " current is the maximum. In the Marconi Instruments TF899 valve millivoltmeter (Fig. 5.21) the natural sensitivity of this type is increased, and operation stabilized, by using it in a bridge circuit, balanced against a similar valve. In this way a bottom range of 0–150 mV is obtained without pre-detector amplification; and the instrument, which has a probe, is effective from 50 c/s to 100 Mc/s.

* For full details see *Wireless World*, November 1953, pp. 516–521.



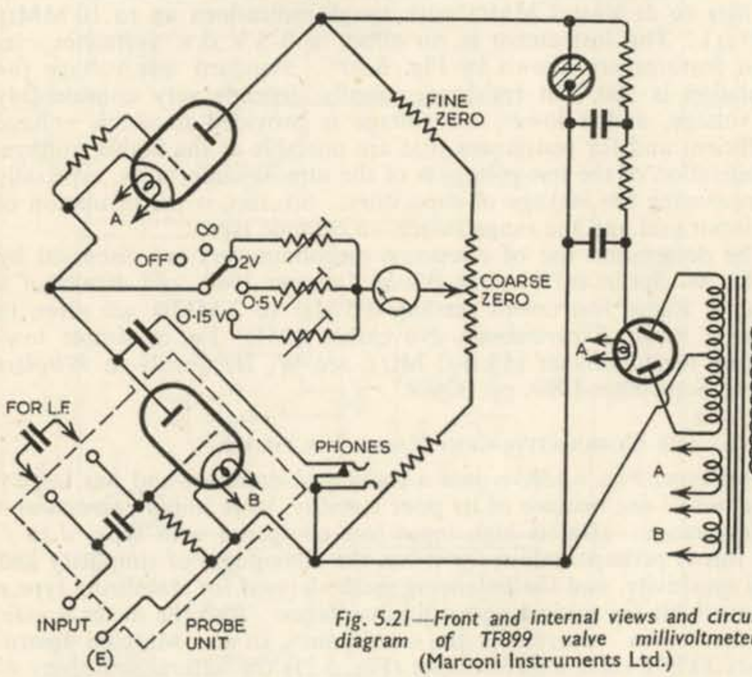


Fig. 5.21—Front and internal views and circuit diagram of TF899 valve millivoltmeter. (Marconi Instruments Ltd.)

5.17. The Anode-bend or Square-law Valve Voltmeter

Although inconvenient in most respects, this type (Fig. 5.13*d*) is the only one that can be used when it is necessary to have r.m.s. readings regardless of waveform. This advantage is obtained along

with a much higher input resistance than types *a–c*, so it is suitable for certain kinds of r.f. measurement. The square law is not obtained automatically, but certain valves, notably small directly-heated output triodes, follow it fairly closely over an appreciable range of a.v. input. The most suitable valve, and initial anode and grid voltages, can be

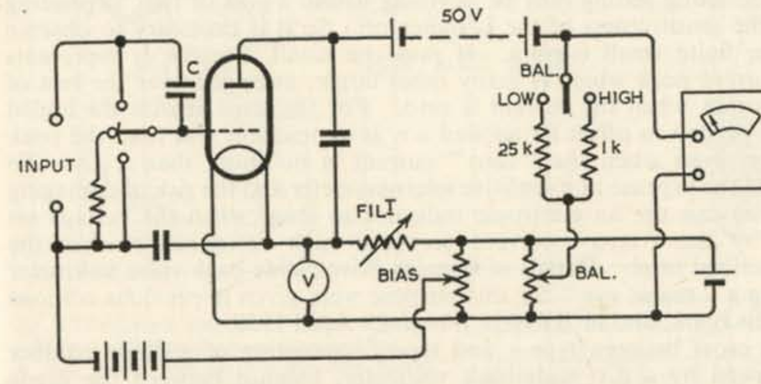


Fig. 5.22—Circuit of anode-bend square-law valve voltmeter for true r.m.s. readings. Provision is made for balancing out the anode current, with or without input

chosen by plotting change in anode current against signal-voltage-squared in each case and noting which gives the longest and best approximation to a straight line. This linearity depends on the resistance in the anode circuit, and is usually best when it is small; so the meter ought not to be loaded up with series resistance, and if more than one range is provided the switching should be such as to keep the resistance constant. For the same reason the anode and grid voltages must be stabilized.

In a typical example (Fig. 5.22) the best initial I_a was found to be 60 μA. The indicator is a Cambridge " Unipivot " meter with ranges of 24, 60, 240, 600, etc. μA; and the initial I_a is balanced out by a tapping across the filament, which is run at about 0·8 normal voltage to minimize risk to the meter. The " High " balance is to enable a deflection to be backed off so that a low meter range can be used to observe a very small change in deflection; this is useful when tuning a circuit precisely to resonance, but must be used cautiously. In fact, this is obviously a purely lab. instrument, not for unskilled use.

C, needed when there is superimposed d.v. or external conduction, must not only be low-loss but also have extremely low leakage.

5.18. The Slide-back Valve Voltmeter

This (Fig. 5.13*f*) is an unpopular type, because it requires a sensitive current indicator as well as a voltmeter, and an adjustment in order to get a reading. It has some advantages, however: low input loss;

almost complete absence of error due to variations in supply voltages and valve characteristics; and ease of discriminating between nearly equal voltages, or setting a voltage to a given level.

Although in theory it may seem simple to adjust the grid bias to the point at which anode current is just reduced to zero, in practice the resulting setting may be anything within a volt or two, depending on the sensitiveness of the I_a indicator. So it is necessary to observe some finite small current. It must be small, because it represents a current peak which is many times larger, averaged over the rest of the cycle, when the current is zero. For the same reason the added bias needed to offset an applied a.v. is appreciably less than the peak value, even when the "zero" current is no more than 1 μA. To avoid the expense of a sensitive microammeter and the risk of damaging it, one can use an electronic indicator to show when the voltage set up by the "zero" current across a high resistance reaches the prescribed level. Details of a mains-driven slide-back valve voltmeter using a "magic eye" for this purpose were given in previous editions of this book, and in *Wireless World*, 28 April 1938.

A cross between type *c* and type *f*, consisting of a diode rectifier followed by a d.v. slide-back voltmeter, balance between the diode output and the variable grid bias being observed by an anode milliammeter which is then switched to read the bias voltage, is described in *Wireless World*, January 1947, by H. W. Baxter. It has the advantage of stable calibration up to 50 V or more with a power source consisting of a 9-V dry battery.

Although not a slide-back type at all, another valve voltmeter that enables much higher voltages to be measured than its own power voltage may perhaps be mentioned here: the Marconi Instruments TF 887A, in which this result is achieved by using a circuit of the diode plus balanced twin triode type, but with the triodes inverted (i.e., input to anodes and output from grids), giving a fractional μ.

5.19. The Reflex Valve Voltmeter

Something approaching automatic slide-back is obtained in this type (Fig. 5.13*e*). It is the "infinite-impedance detector" with the addition of a meter to read the cathode current. It is also akin to the cathode follower, but its negative feedback is d.v. only, as the load is by-passed for a.v. by C. The sensitivity, in meter-current change per signal volt, is rather low, because the scale is, as it were, telescoped by the sliding bias; but this very fact gives it the advantage of a much longer voltage range than types *b* and *d*. And it lends itself to multi-ranging, by varying R. However, the valve-supply voltage must be adequate: of the order of three times the highest peak-signal voltage. The scale is almost perfectly linear above about 2 V. The readings are approximately proportional to peak values for large voltages, 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

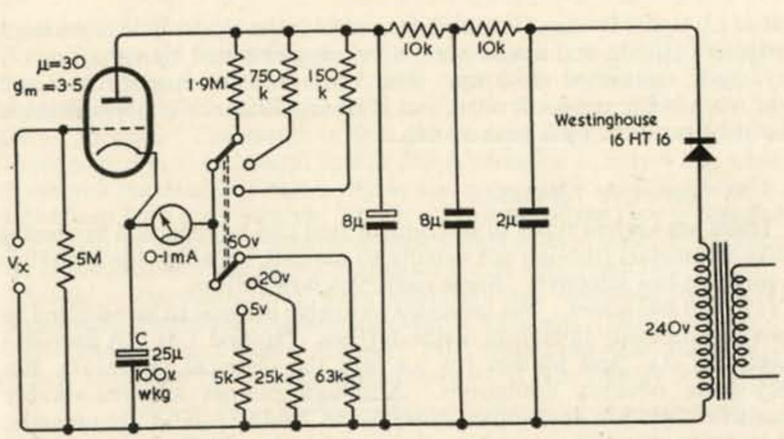


Fig. 5.23—Circuit diagram of reflex valve voltmeter with supply-voltage compensation. It is a useful and easily-made type for general purposes

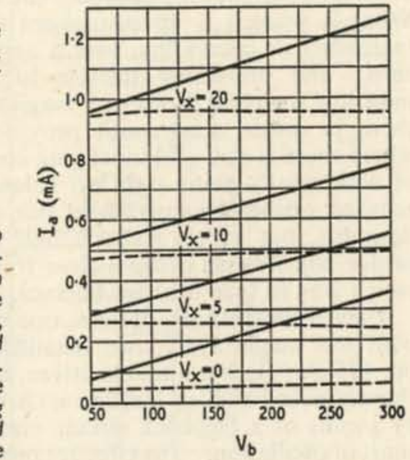


Fig. 5.24—The full lines, which apply to a reflex valve voltmeter without compensation, show how the meter deflection depends on the supply voltage, V_b. The readings were taken with R = 55 kΩ, μ = 30, g_m (nominal) = 3·5 mA/V, and C = 50 μF; and V_x is specified in r.m.s. values of 50-c/s sinusoidal voltage. The dotted lines show the effect of adding compensating resistance (= μ times load resistance) from anode to load resistance

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 (*Wireless World*, October 1949, pp. 401–4) 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. 7.20). Fig. 5.23 shows a practical three-range circuit using a 0–1 mA meter, incorporating this modification, which renders supply-voltage stabilization unnecessary. In Fig. 5.24 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, but if its conductance is objectionable it can be replaced by a push switch.

5.20. 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. 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. 5.33) 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 ", are detailed in *Wireless World*, May 1951, pp. 175–8. 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. 4.17, set just short of oscillation. The effective range of signal is 10 μV–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, described in *Muirhead Technique*,

* " 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.*, October 1939, pp. 649–655.

April 1950, is of this type, which normally has a bandwidth proportional to the 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 unsuitable if the signal frequency is not very constant.

Selective amplifiers and wave analysers are discussed by G. H. Hickling in *Electronic Engineering*, March 1952, pp. 120–6.

(3) *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.

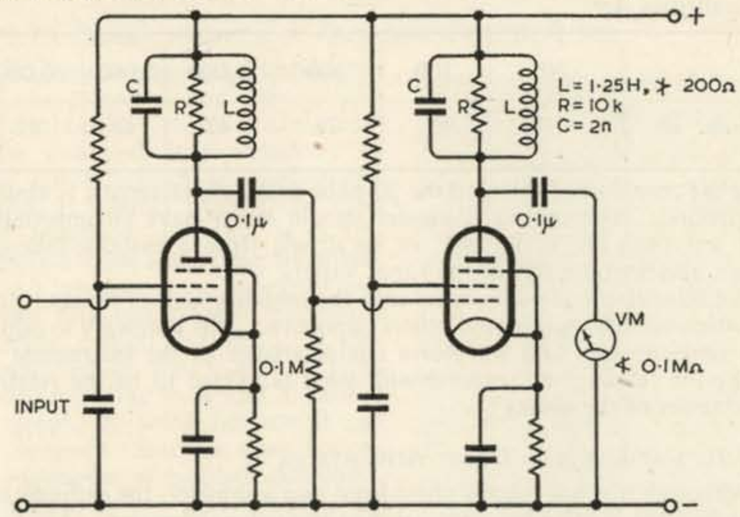


Fig. 5.25—Circuit of a " weighted " amplifier to compensate for the characteristics of the average ear (Fig. 14.46) to indicate loudness of hum or other low-level sound

amplifier to have a frequency characteristic similar to that of the ear (Fig. 14.46). 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. 5.25 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. 14.46 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 $1\frac{1}{4}$ 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. 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.*

5.21. 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 all the other indicating instruments that have been described 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 the radio investigator.

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

* "Electrical Noise", by D. Maurice, G. F. Newell, and J. G. Spencer. *Wireless Engineer*, January 1950, pp. 3–12.

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. The idea that nothing much can be done without an instrument costing £50 or more is erroneous. While these ready-made and versatile oscilloscopes are very convenient, much useful work can be done with a cheap type of 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 be pointed out in connection with the measurement of distortion) not even the most satisfactory technique.

5.22. 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—not that it is impossible to use some types for either.

All laboratory c.r. tubes are provided with two pairs of plates for electric deflection (Fig. 5.26), distinguished as the X and Y pairs (corresponding to X and Y axes of 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. 5.27*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

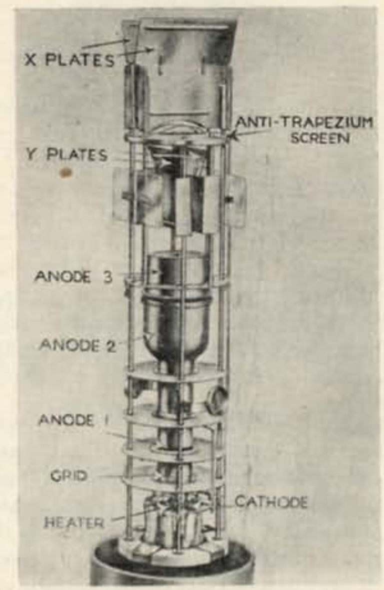


Fig. 5.26—Typical electrode construction of single-beam cathode-ray tube. (A. C. Cossor Ltd.)

* They are introduced in the author's *Foundations of Wireless* (5th ed.), and the following books are recommended for clear simple treatment of the subject:
Cathode-ray Oscillographs (4th ed.), by J. H. Reyner (Pitman, 1951).
The Cathode-ray Tube Handbook, by S. K. Lewer (Pitman, 1947).

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 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

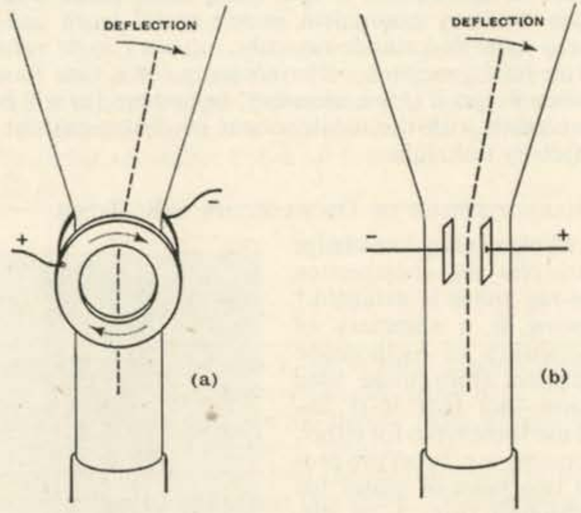


Fig. 5.27—Showing how (a) *magnetic deflection and* (b) *electric deflection are accomplished, and the relationship between polarity of current or voltage and the direction of deflection*

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. The main object of using such high voltages is for photographing transient deflections. For more general purposes one should use the lowest anode voltage that gives adequate focusing and brilliance, for this economizes in tube life, power unit, and signal amplification. An anode voltage over 2,000 is exceptional for oscilloscopes; 1,000–1,500 V is usual; and reasonably good results can be obtained with 500–1,000, or even less with gas-focused tubes.

The usefulness of an oscilloscope " picture " or trace depends not so much on its size as on the ratio of size to spot diameter, so fineness of focus is important. Gas-focused tubes, in which there is only a partial vacuum, and focus is obtained at a critical beam current

adjusted by grid bias, have gone almost entirely out of use, which seems a pity, for they give good results (for deflection frequencies of up to a few hundred kc/s) at only a few hundred volts.

High-vacuum 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 general the focus improves with increasing final anode voltage, which tends to compensate for the decreasing deflection sensitivity. 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. 5.26, the theory of which is explained by B. C. Fleming-Williams in *Wireless Engineer*, February 1940, pp. 61–4.

If the deflection sensitivity is, for example, 400/V mm per volt, and the final anode is run at 800 V, 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 the deflection sensitivities of X and Y plates are generally not the same. 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.

An important characteristic, more especially at high frequencies, is the signal-input impedance. 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.

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.

5.23. Double-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 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 Cossor double-beam tube, which enables this to be done, is particularly recommendable, as it costs no more than the corresponding single-beam type, with which it is interchangeable. The device for producing the second, independently-deflected, beam is simplicity itself—one extra plate situated midway between the first pair of deflector plates that the cathode ray encounters (the Y plates). As at this stage the ray is undeflected and is in a diffuse condition it is split in halves by this central plate, which is internally connected to the main anode and therefore at zero or reference potential so far as the deflection system is concerned. This extra plate can be seen in Fig. 5.28.

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. Since the horizontal displacement is the same 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!

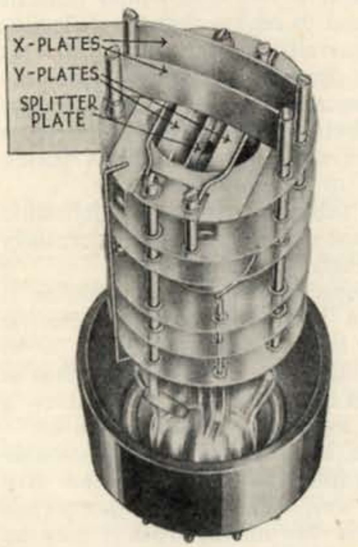


Fig. 5.28—Deflection plates of double-beam cathode-ray tube. (A. C. Cossor Ltd.)

The method of use is almost self-explanatory, as, apart from the separate action of the two Y plates, everything is the same as for the conventional single-beam tube. If desired, one picture can be raised clear of the other on the screen by applying a suitable steady voltage in series with the signal voltage.

In some tubes, multiple beams are obtained by means of a corresponding number of electron guns.

For example, a type made by Twentieth Century Electronics contains four separate guns.

To obtain the effect of more than one beam, use is sometimes made of an " electronic switch " to hand the Y plates over in rapid succession to the several signal sources. Each trace then consists of dotted lines.

5.24. 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 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. 9.9). By applying the input and output signals of an amplifier to X and Y plates, it is possible to compare amplitude, phase, and also *waveform*; for 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 very many 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.

5.25. Power Supplies

Coming now to auxiliary apparatus, the one essential thing is the power source. The heaters of c.r. tubes are similar to those in valves, but as it is usual for the anode to be earthed the cathode is at full negative h.t. 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.

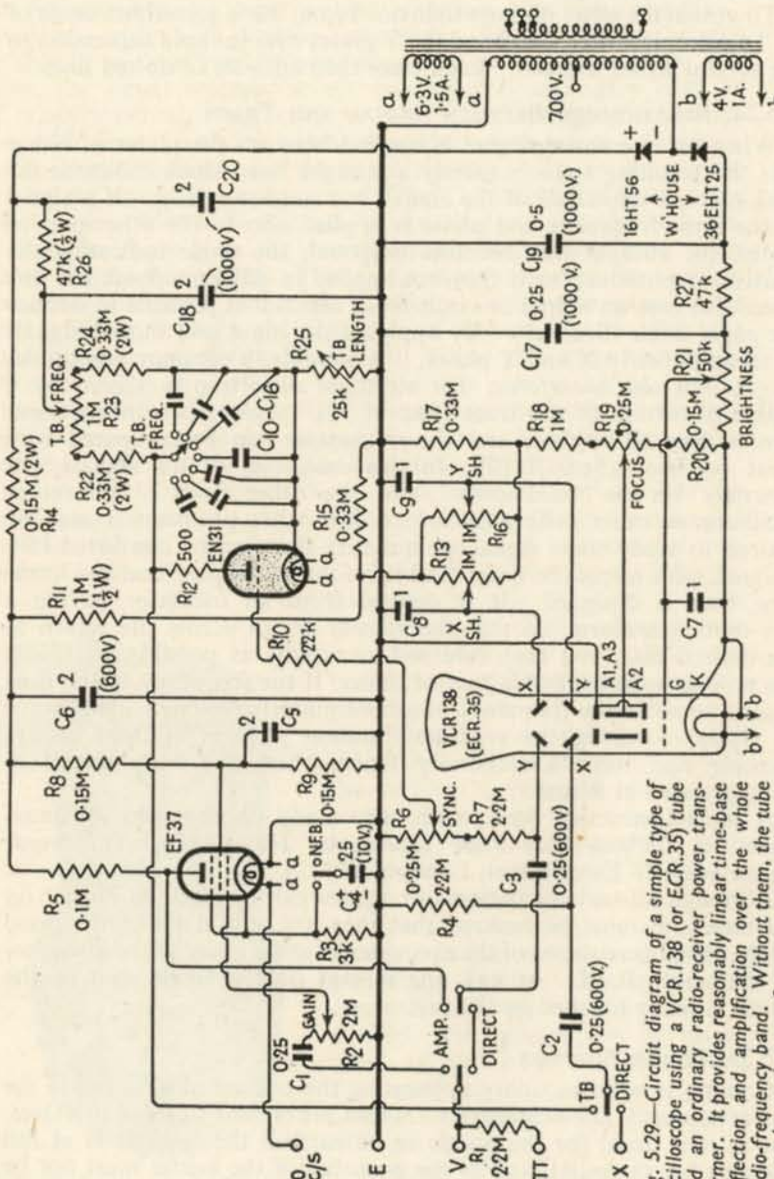


Fig. 5.29—Circuit diagram of a simple type of oscilloscope using a VCR.138 (or ECR.35) tube and an ordinary radio-receiver power transformer. It provides reasonably linear time-base deflection and amplification over the whole audio-frequency band. Without them, the tube can be used at any frequency up to v.h.f.

Provision of the final-anode voltage is simplified by the trifling current demand. The tube itself takes an almost negligible anode current, of the order of 20 μA, and the total loading is determined mainly by the potential divider used to obtain focusing potentials. Gas-focused tubes have only one anode, the focus being adjusted by heater current and grid bias, so it is not essential to have a potential divider; but even with high-vacuum tubes, where focusing anodes have to be supplied at reduced voltages, the current drain need not be more than about 1 mA 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 Fig. 5.29, which is the circuit of a simple oscilloscope described by the author in *Wireless World*, March 1950 (and 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. The negative-voltage circuits are shown below the heavy horizontal earth line. The potential divider is made up of R_{17}–R_{21}, with R_{27} for smoothing. Beam current, which determines brightness, is controlled by grid bias (R_{21}), and focus is controlled by the potential of the second anode, A2 (R_{19}). In most tubes its voltage is somewhere about one-fifth that of the final anode. The tube shown has the first anode A1 joined internally to A3, but this is not so with all types.

The final anode, often called just " the anode ", is usually earthed, and the potentials of the deflection plates are reckoned relative to it. All the plates must at all times have conducting paths between them and the anode, such as R_4 and R_7 in Fig. 5.29, otherwise the trace is free 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 R_{13} and R_{16}. To enable the shifts to work both ways from centre, the negative potential of the ends joined to R_{17} is matched by a positive potential from across R_{15}. Double-beam tubes need separate Y shifts for the Y plates. In more refined oscilloscopes, such as the Cossor 1035, 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. An alternative scheme is to use sources of known and variable voltage and frequency with which to compare the signal. The construction of a voltage-calibrating source is described by W. Tusting in *Wireless World*, August 1951; and a commercial type in Sec. 4.20.

For the sake of portability most oscilloscopes comprise the c.r. tube, its power unit, and usually a time-base generator and signal amplifiers, all built into one box. But for bench use there are some advantages in supporting the tube by itself in a retort clamp (Fig. 5.30). This enables it to be set at the most convenient angle and its deflection plates connected to the test points more directly and with less stray capacitance than is possible with the conventional oscilloscope. The power unit, being connected by a flexible cable, can be kept

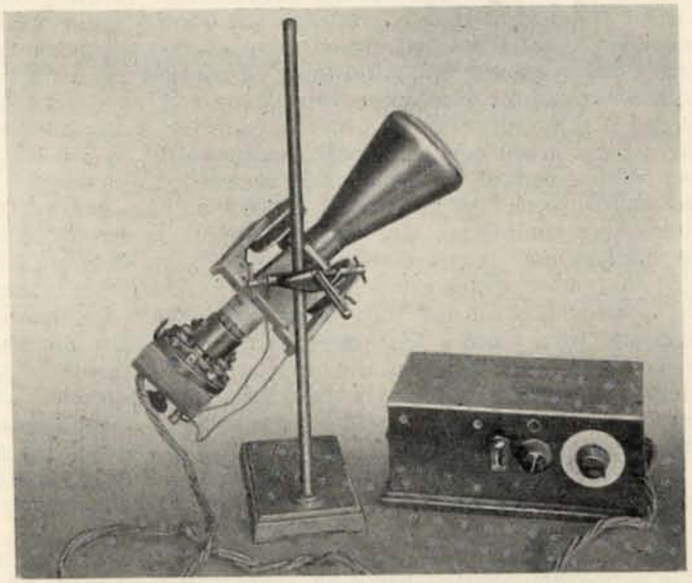


Fig. 5.30—A chemical retort stand is a convenient mounting for a cathode-ray tube, enabling the screen to be adjusted to a convenient angle for viewing, and the deflector-plate terminals to be brought close to the apparatus being tested. Note the 50-c/s deflector coils around the neck, with amplitude control at the base of the tube

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 transformers are allowed 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 in *Wireless World*, December 1951.

5.26. The Sinusoidal Time Base

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 quite often it is 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. 5.30, 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. 5.31, but there is no need to keep strictly to these.

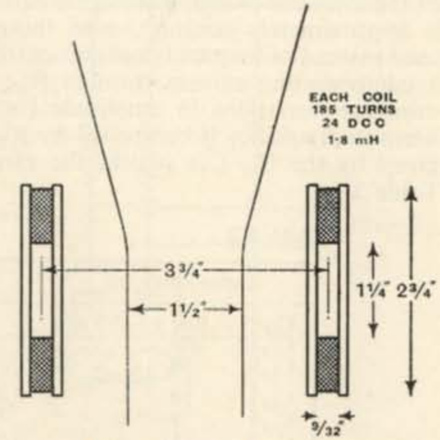


Fig. 5.31—Dimensions of coils suitable for providing a uniform 50-c/s deflecting field

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 reliable, 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 300 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.

5.27. Linear Time-base Systems

However, sometimes one wants the whole of a signal to be reproduced on a linear time base, and for this purpose a great variety of circuits have been worked out, some using gas-discharge valves and others depending only on ordinary " hard " (i.e., high-vacuum) valves. For details of these many varieties O. S. Puckle's *Time Bases* (Chapman & Hall) should be consulted. The basic principle is to charge a capacitor through some constant-current device, such as a high-impedance pentode, and then when it reaches a certain voltage to discharge it suddenly. The voltage across it therefore consists alternately of periods of uniform increase and practically instantaneous decrease, which is what is wanted for the X plates. By altering the capacitance of the capacitor, or the anode current of the valve, or both, the frequency of the saw-tooth wave can be varied as desired. Gas-discharge (" soft ") valves are capable of passing a very heavy current for rapid discharge of the capacitor. Only one such valve is required, in addition to a constant-current valve, if used. In Fig. 5.29, the EN31 is the discharging valve, and as only about one-sixth of the available charging voltage is actually needed, the rate of charging is approximately constant, even though ordinary resistance (R_{23}) is used instead of a special constant-current device. R_{24} is there to pass a compensating current through R_{25} (controlling the bias) so as to minimize variations in amplitude (or length) of the time deflection when its frequency is controlled by R_{23}. Step-control of frequency is given by the C_{10}–C_{16} switch, the ranges being approximately as in Table 5.2.

Table 5.2

	Capacitance, nF	*Frequency, c/s*
C_{10}	300 (= 0·3 μF)	12–53
C_{11}	100	35–160
C_{12}	30	120–530
C_{13}	10	350–1,600
C_{14}	3	1,200–5,000
C_{15}	1	3,600–14,000
C_{16}	0·3	13,000–40,000

R_6 is to enable a small controllable fraction of the " work "—the Y signal—to be impressed on the grid of the discharge tube for synchronizing the time sweeps so as to keep an exact whole number of work cycles on the screen.

There are various methods of making the sweep more linear, other than the device of charging through a pentode. Where a push-pull output is used—and it is strongly recommended—a very convenient and effective method is the one shown in Fig. 5.32. The linearity control is adjusted until a number of cycles of a Y signal are all equally spaced. The output stage has normal component values for that type (sometimes called the " long-tailed pair "). Note the convenient method of X shift.

If much higher sweep frequencies are required, a soft valve is unsuitable. Hard-valve time bases are generally more elaborate, but more stable in operation and can be used up to 1 Mc/s or more. The best-known is the Puckle type, one form of which is shown in Fig. 5.33. C is charged through V1, and the voltage across it rises at a uniform rate. V2 is connected in parallel with C, but as its grid is initially biased negative by the drop across the amplitude control it passes no current until the voltage across C reaches a predetermined amount. Once a small anode current flows through V2, it causes a signal to be passed to V3, which makes the grid of V2 more positive, thereby enormously accelerating the discharge. When discharge has taken place the valves return to their original condition for the next stroke. The control grid of V3 is used for synchronization.

An ingenious variation of the Puckle time base, in which the synchronizing system not only pulls the time-base frequency into exact integral relationship with the work, but automatically controls

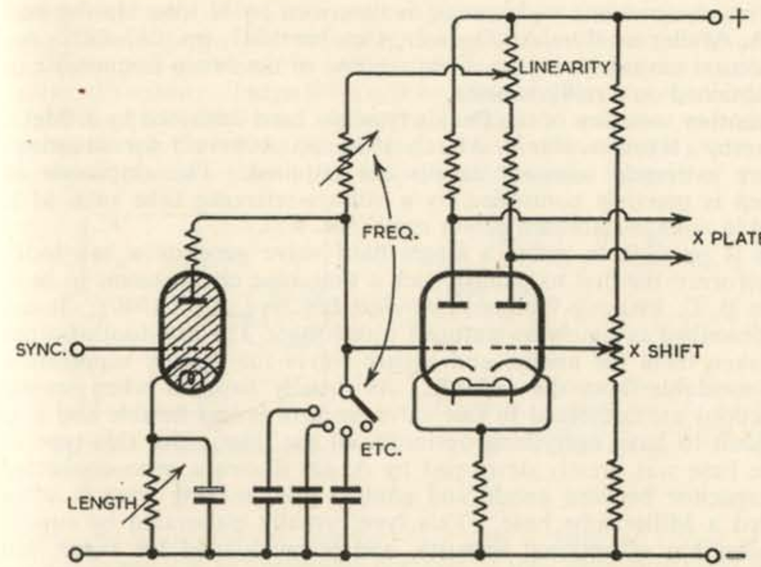


Fig. 5.32—This symmetrical amplifier is eminently suitable for either X- or Y-plate deflection. Here it is shown combined with a soft-valve time-base generator, to which it contributes a linearizing adjustment

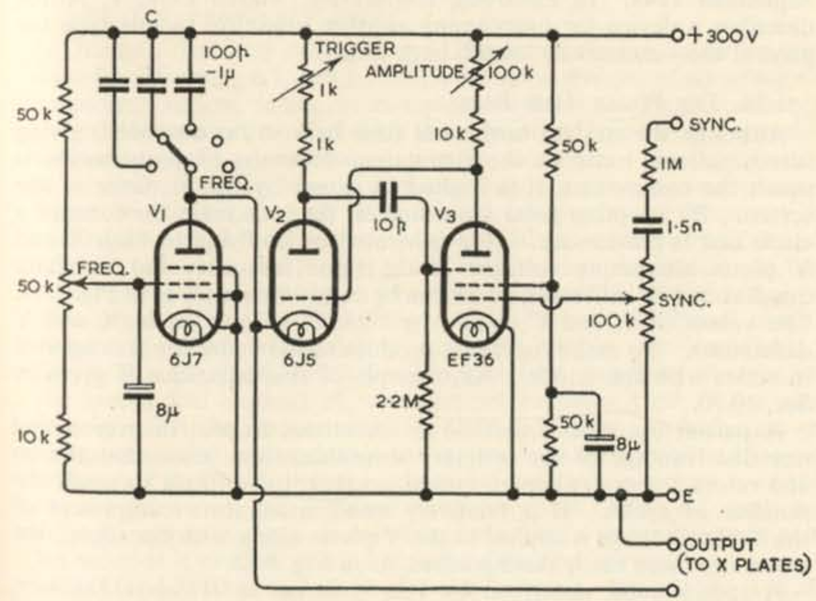


Fig. 5.33—Circuit diagram of Puckle hard-valve time-base generator

its frequency over a wide range, is described by H. den Hartog and F. A. Muller in *Wireless Engineer*, October 1947, pp. 287–292. An incidental advantage is that direct reading of the sweep frequency can be obtained on a milliammeter.

Another variation of the Puckle type has been described by J. McG. Sowerby (*Wireless World*, March 1950, pp. 108–110) for situations where extremely constant results are required. The amplitude of sweep is precisely controlled by a voltage-reference tube such as is used in voltage-stabilized power units (Sec. 4.3).

It is possible to make a single hard valve generate a saw-tooth waveform; the first to publish such a time-base circuit seems to have been B. C. Fleming-Williams (*Wireless Engineer*, April 1940). It can be described as simply an untuned transitron. The saw-tooth output is taken from the anode, and square waves for flyback suppression are available from the cathode. As usually happens when several functions are combined in one valve, control is less flexible and it is difficult to have everything optimum all the time. But this type of time base was greatly developed by A. D. Blumlein, who connected a capacitor between anode and control grid, making what is often called a Miller time base. This type, usually elaborated by sundry diodes, has exceptional linearity, and is much used for radar, but is less suitable for oscilloscopes because it is more troublesome to control the frequency over a wide range. A good account of it, with practical circuits, is given by B. H. Briggs in *Electronic Engineering*, September 1948. In *Electronic Engineering*, March 1948, V. Attree describes a device for overcoming another objection to this type for general use—excessively long flyback time.

5.28. 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 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. 5.34. The values of 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. 5.35.

A special tube, described by von Ardenne in *Wireless Engineer*, January 1937, enables a spiral time base 160 in long to be displayed

on a 4-in screen! Similar results can be contrived with ordinary tubes by modulating the a.v. shown in Fig. 5.34 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 instance, can be shown instantly on the screen instead of tracing

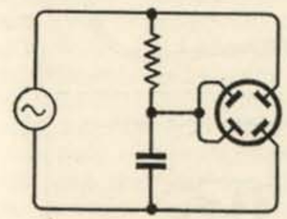


Fig. 5.34—Phase-splitting circuit for generating a circular time base

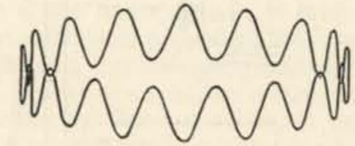


Fig. 5.35—Signal display on sinusoidal time base, "opened out" by addition to signal voltage of 90° displaced time-base voltage

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. 9.35.

5.29. 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. The X deflection is generally obtained from a perfectly standard time-base generator—linear or sinusoidal—and is translated into terms of frequency by making it control the frequency of the signal generator proportionately to deflection. So it does not much matter what the law of the base is as regards time, so long as frequency is controlled to the same law. This rapid variation of frequency to and fro over a selected band is commonly called wobbulation, and the means are described in Sec. 4.29. Fig. 5.36, which is a diagram of a complete set-up, using the mechanical method of wobbulation, explains itself. Care must be taken that the frequency of the time base is not so high that the apparatus under test is unable to respond quickly enough. Using an ordinary "fast" c.r.t. screen, about 10 sweeps per second is needed for the eye to see the response reasonably well; with a.f. tests, and r.f. sets containing very sharply-tuned circuits, a sweep time of one to ten seconds is needed, and a long-afterglow c.r.t. screen.

In Fig. 5.36 the frequency of sweep is set by the motor, and the time-base generator is pulled into step. The general connections of

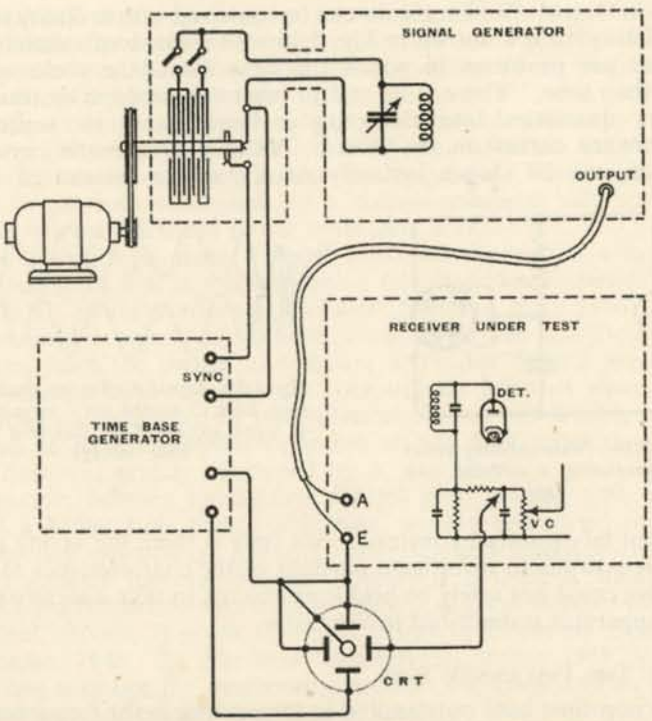


Fig. 5.36—Complete system for continuous observation of receiver response characteristics, using mechanical frequency modulation of signal generator by motor-driven capacitor, synchronized with linear time base. The output of the receiver under test is applied to the other pair of deflectors. In the symbol for the cathode-ray tube the central spot represents the anode, which is normally connected to the " earthy " plate in each pair

a system using electronic wobbulation (Sec. 4.29) are similar, but the t.b. generator sets the pace and its output varies the frequency of the signal generator. A reprint published by *Wireless World* comprises the author's designs for a simple oscilloscope and for a wobbulator, showing the interconnections.

5.30. Deflection Amplifiers

It may often happen that a signal under investigation is too small to produce a useful deflection, more especially with high-voltage tubes, without preliminary amplification. This is rather inconvenient and one of the best reasons for keeping the tube voltage to a minimum. The inconvenience is not only that amplifiers represent so much more additional gear to provide, but also so much more to think about when considering the circumstances of the experiment. The amplifiers must give the desired amplification, which must usually be a known

quantity, and certainly must not vary with signal frequency or amplitude, or with time, or any other irrelevant quantity.

The frequency band over which a resistance-coupled amplifier gives constant gain is limited at the bottom end by the series capacitance of blocking capacitors and at the top end by stray shunt capacitance. The former is easier to cope with, for it is just a matter of giving the blocking capacitors and their associated resistors a sufficiently long time constant (Sec. 14.15), but it must be remembered that a 20-c/s square wave is perceptibly distorted if the time constant is even as long as 1 sec (= 1 MΩ-μF). Some deflection amplifiers avoid the lower limit entirely by using direct coupling throughout. Raising the upper frequency limit is more difficult, for stray capacitance cannot be reduced indefinitely. When all possible has been done about that, one can only reduce its effect by reducing the resistance across which it acts, and that has the disadvantage of reducing the gain of the stage, severely limiting the signal amplitude and increasing the tendency to distortion. It is futile to use c.r. equipment to study

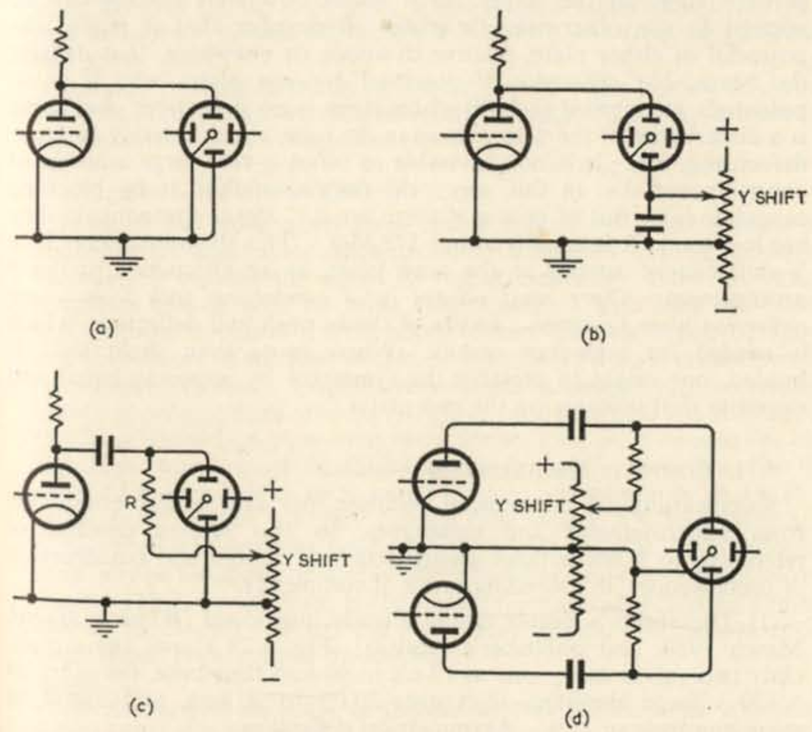


Fig. 5.37—Four alternative methods of connecting Y plates: a is direct connection, modified in b to provide controllable shift voltage. In c, the anode-supply component of voltage is removed; and in d there is symmetrical deflection with controllable shift

distortion if the apparatus is putting in distortion of its own. The top frequency limit can be built out to several Mc/s by judicious use of inductance (Sec. 14.16), and the use of considerable power. A usual practice in oscilloscopes is to provide high-gain low-distortion amplification for a.f. and perhaps low r.f.; and low-gain amplification, unavoidably limited in undistorted output, up to several Mc/s. To prevent the capacitance of the c.r.t. plates and connections affecting the amplifier bandwidth, cathode followers are sometimes used as output stages. And, as already explained, push-pull output is most desirable, especially with c.r. tubes that are not designed for asymmetrical deflection.

It is quite easy to get muddled about connecting the signal to the c.r.t., with resulting undesired short circuits and initial deflections, especially when working from amplifiers. Fig. 5.37*a* shows direct asymmetrical connection; one plate of each pair is kept at constant potential, the same as the c.r.t. anode, and the other receives the signal. But the positive bias of the valve anode deflects the spot, perhaps right off the screen, so *b* shows how shift voltage can be applied to the otherwise idle plate. Remember that it is not the potential of either plate, relative to anode or elsewhere, that deflects the beam, but *difference* of potential *between* plates. So if equal potentials are applied to both plates there is no deflection; but there is a disturbance of the field pattern in the tube, which is likely to cause defocusing, etc. It is not advisable to offset a very large amount of superimposed d.v. in this way; the normal method is by blocking capacitor (*c*). But of course if there are d.v. *signal* components they are lost too. R is usually about 1–2 MΩ. This diagram shows how Y-shift can be applied to the same plate, as an alternative to the *b* arrangement. *There must always be a conducting path from every deflection plate to anode.* Lastly, *d* shows push-pull deflection, which is needed for high-class results. Where more than slight shift is needed, one ought to preserve the symmetry by imposing equal and opposite shift voltages on the two plates.

5.31. Complete Oscilloscopes

Specifications of commercial oscilloscopes are readily obtainable from advertisements and catalogues, so this section consists of references to a few articles giving details of design and construction of oscilloscopes, in ascending order of complexity.

(1) The author's simple design already mentioned (*Wireless World*, March 1950, and publisher's reprint); Fig. 5.29 shows the circuit. Only two valves used; one as 12-c/s to 40-kc/s time base, the other as ×120 voltage amplifier, level over 20 c/s to 20 kc/s, and useful to some hundreds of kc/s. Asymmetrical deflection.

(2) " Television Oscilloscope ", by W. Tusting (*Wireless World*, June 1952, pp. 233–6, and July 1952, pp. 281–3). Three-valve time base, similar to Fig. 5.32, giving linear and symmetrical deflection

16–10,125 c/s; two-stage Y amplifier, also with push-pull direct-coupled output to plates, gain ×300, and bandwidth about 4 c/s to 0·6 Mc/s; calibrator giving 50 mV to 50 V p-p at 50 c/s. (*Correction:* R_{16} should be 40 kΩ instead of 150 kΩ.)

(3) " General-purpose Oscilloscope " (modified from radar unit), by J. F. O. Vaughan (*Wireless World*, May 1948, pp. 161–5, and reprint). Valve line-up and frequency ranges somewhat similar to (2), but Blumlein time base with separate buffer stage for synchronizing, and amplifier low-frequency cut-off higher.

(4) " Cathode-ray Oscilloscope ", by S. A. Knight (*Wireless World*, December 1948, pp. 432–6, and reprint). Three-valve Puckle time base, 5 c/s to 200 kc/s; single-valve Y amplifier, flat 60 c/s to 0·55 Mc/s; two-valve wobbulator (optional), 465 kc/s ± 12 kc/s.

(5) " Universal Oscillograph ", by G. L. Hamburger (*Electronic Engineering*, January 1947, pp. 7–10 and 22, and February 1947, pp. 51–57). Puckle time base, with linearized and symmetrical deflection, 0·5 c/s to 275 kc/s and single-stroke working; elliptical time base, using external a.c. up to 60 kc/s; single-stage Y amplifier with variable gain and level; voltage calibrator, 0·1–200 V; and frequency calibration up to 10 kc/s.

Some interesting information is given in the same journal (October 1947) by H. L. Mansford on a waveform monitor: a c.r. equipment for the measurement of voltage and time.

5.32. Photographing Oscillograms

Most of the work apart from rather specialized investigations on atmospherics and other uncontrollable transients can be observed visually, by causing the signal to repeat cyclically. Slow individual transients can be seen by using special screens with long afterglow. But for rapid transients it is necessary to have recourse to photography, and the makers of the tube should be consulted for their recommendations regarding this. Nothing very special is necessary, however, for taking photographs of stationary figures on the screen. It is essential that they should be *quite* stationary for as long as is needed for the exposure, which must be found by experiment. Something of the order of a second may be required for a low-voltage tube and a stop of about *f*6. With a higher voltage tube, Super XX Pan. film, and *f*2 aperture, exposures as brief as $\frac{1}{50}$th sec are possible. An exposure must not be *too* rapid or it may not include a complete cycle of movement. (This, by the way, is the basis of a method of testing camera shutter accuracy.) Ilford, Ltd., of Ilford, Essex, will advise about oscillogram photography on request.

References:

The Photographic Recording of Cathode-ray Tube Traces, by R. J. Hercock (Ilford Ltd.) (with bibliography).

" Photography of Cathode-ray Tube Traces ", by H. F. Folkerts and P. A. Richards. *R.C.A. Review*, October 1941, pp. 234–244.

" Photographing C-R Tube Screen Traces ", by T. A. Rogers and B. L. Robertson. *Electronics*, July 1939, p. 19.

5.33. 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 by the simple

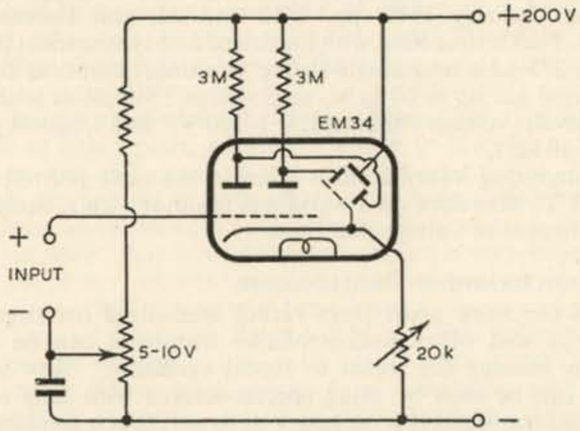


Fig. 5.38—Method of using the EM34 as a voltage indicator with controllable feedback to increase its sensitivity

expedient of connecting in series with the cathode a resistor of a few kilohms, preferably variable. This introduces positive feedback; 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. 5.38 shows the makers' recommendation.*

Tuning indicators are popular as balance indicators in bridges (for example, Sec. 7.12), and are useful for indicating maximum in resonance methods. In slide-back methods they may even be used as substitutes for meters.

The Mullard DM70, being a sub-miniature measuring only $1\frac{1}{2}$ in × $\frac{3}{8}$ in, is of especial interest where space is limited.

Ordinary neon lamps are most useful in the laboratory as indicators and voltage stabilizers. It should be known that they are sold with or

* *Wireless World*, April 1947, p. 150.

without a resistance in the cap: for laboratory purposes they are more useful without the resistance, but as the current then rises almost 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.

5.34. 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 loudspeakers.

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 the 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 in circulation for some time 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 in *Wireless Engineer*, March 1943, pp. 109–126, 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).

For the use of standards and sub-standards for calibration, see " A R.A.F. Calibration Centre ", by W. H. Ward and others, in *Proc. I.E.E.*, Pt. III, January 1950, pp. 49–55.

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 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 G and B. The reducing of more 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 subject 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 vast number of different 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. 6.1*a*, 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

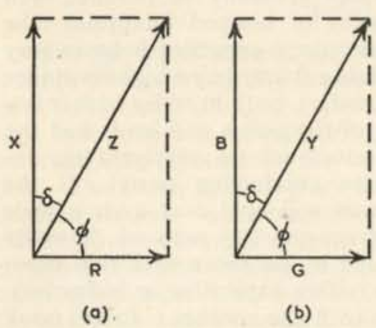


Fig. 6.1—Vectorial representation of (a) resistance R, reactance X, and impedance Z, and (b) conductance G, susceptance B, and admittance Y, to illustrate the numerous alternative methods of specifying their phase relationship

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:

$$\text{(i)}\quad \cos\phi = \frac{R}{Z} = \frac{G}{Y} = \sin\delta = \text{power factor}$$

$$\text{(ii)}\quad \cot\phi = \frac{R}{X} = \frac{G}{B} = \tan\delta = \frac{1}{Q} = \text{dissipation factor (or loss tangent)}$$

$$\text{(iii)}\quad \tan\phi = \frac{X}{R} = \frac{B}{G} = \cot\delta = Q = \text{magnification factor}$$

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. In a resistor that is supposed to be non-reactive it should 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 (0–10). 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

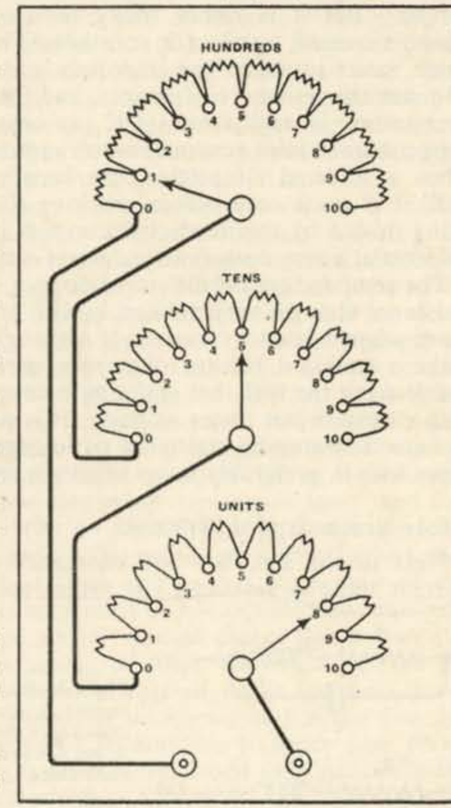


Fig. 6.2—Connections of a triple-decade resistance box, providing 1,110 Ω in steps of 1 Ω

0·1–0·2 per cent and is reliable at all audio frequencies and well into the radio frequencies. In modern practice, switch contacts are kept inside the box, with only a pointer knob external. To preserve these valuable instruments from damage it is worth making a table of maximum permissible current for each range, and the 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. "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 per cent 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. 14.39). 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 cent 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.

The temperature coefficient of copper is about 0·4 per cent, 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

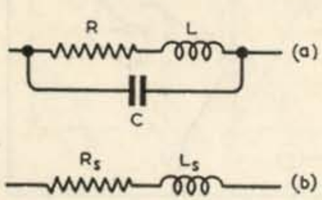


Fig. 6.3—Diagram a represents with good approximation a resistor, L and C being the residual inductance and capacitance. At any one frequency, L and C can be represented as an equivalent series inductance L_s, as at b; but the resistance R_s is in general different from R

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 per cent high at 1 Mc/s. For minimum 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 10–1,000 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 often 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 sytem is 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

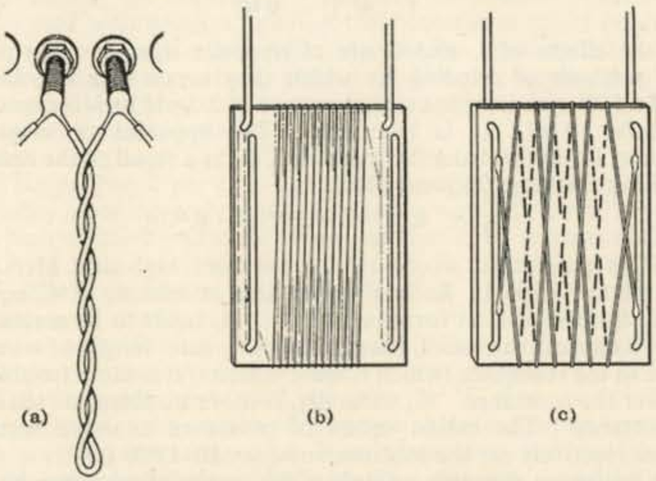


Fig. 6.4—Three of the many methods of minimizing reactance in resistance elements. The "twisted-hairpin" type (a) is suitable for low resistance; b consists of fine wire wound on a thin mica card, which is stiffened by the thick wire connecting leads. In c (the Ayrton-Perry method) there are two windings in parallel, in opposite rotation to neutralize self-inductance

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 films; and although they do not yet compare favourably in precision with the best wire-wound standards, 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. 4.24 shows an example of use in a calibrated attenuator up to over 300 Mc/s. But whatever makers may say, it is probably wise to check them rather more frequently than wire-wound types.

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 1–100 Ω, are usually made of short

lengths of wire sufficiently thin to keep the skin effect within tolerable limits. The gauges for a 1 per cent rise are given in Table 14.8.

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.

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 14.24), 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 wound with a thicker gauge.

With low values it is important to consider the resistance of the switch contacts and leads. The method of use, wherever possible,

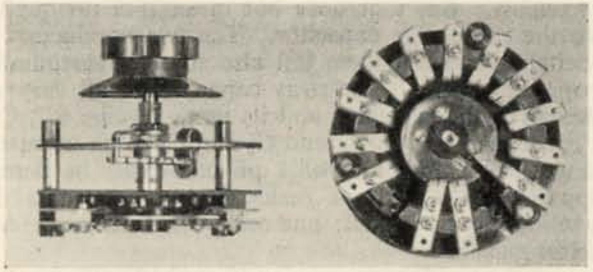


Fig. 6.5—Two views of precision type of switch designed to provide a wide range of switching combinations with low contact resistance. (W. G. Pye & Co. Ltd.)

should eliminate these from the calculation, by working in differences as explained in Sec. 3.9. But even so, these undesired resistances should be minimized; in particular, a reliable switch should be used. Fig. 6.5 shows one specially designed for instrument work, available with a variety of poles and ways, with contact resistances of 0·3 or 1 mΩ. For less precise purposes the comparatively cheap wafer types used in receivers are generally satisfactory.

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

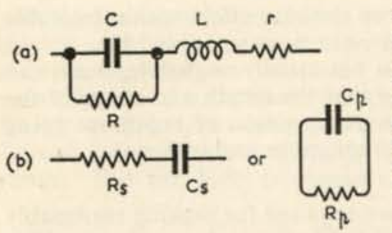


Fig. 6.6—To represent the behaviour of a capacitor over a range of frequency it is necessary to include parallel as well as series resistance (a). At any one frequency, resistance and reactance can be represented either in series or parallel (b). Except at such high frequency that the reactance of L 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. 6.6*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. 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. 6.7, 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. 9.15) all capacitances to earth are lumped in with the measuring apparatus, which reads only

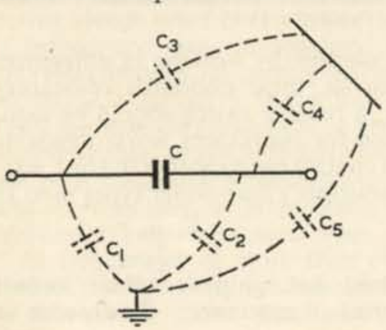


Fig. 6.7—The stray capacitances of an unscreened capacitor can be represented as shown, C_1 and C_2 being to earth and C_3 and C_4 to non-earthed surroundings. By the use of screening, these variable capacitances can be reduced to one fixed capacitance included in the calibration

what is called the direct capacitance. In Fig. 6.7 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 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? As far as possible, connecting the capacitor should introduce only its own capacitance from its terminal inwards; but it is difficult to ensure this. 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 they are very useful, if one can afford them.

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 backlash. 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 limited to ordinary " commercial " types. For information on the technique of making direct-reading scales to an accuracy of 0·01 per cent, see W. H. F. Griffiths, *Wireless Engineer*, November and December 1943. A recent Sullivan development is a capacitor continuously variable over 100 pF (small enough to give a very open scale) and associated with ten steps 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. 6.8) 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.

Such instruments are not likely to be within the means of small laboratories. The Mullard Type F is a good compromise, with a laboratory specification at moderate cost. 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. 6.9). Backlash or whip anywhere between the vanes and the scale must of course be negligible. A good geared drive is a great asset. Fig. 6.10 shows a two-speed type that gives quick adjustment to any part of a clear 1,000-line scale. As regards vane shape, although

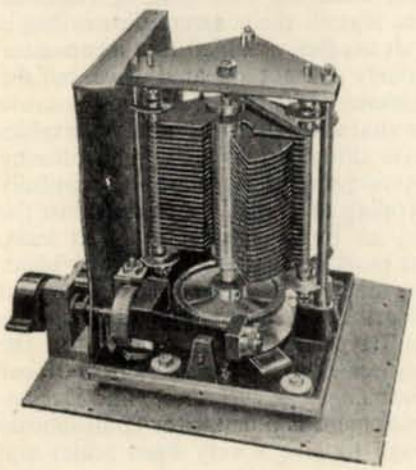


Fig. 6.8—Precision laboratory capacitor (Type D-14) designed to have an extremely small loss angle. The inner screen (shown mostly cut away) is connected to the rotor or low-potential terminal; there is a separate earthed screen, not shown except for the base plate. Note the earthed guard disks mounted between quartz insulators supporting the stator or high-potential system. (Muirhead & Co. Ltd.)

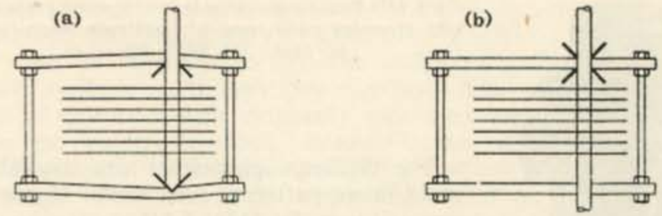
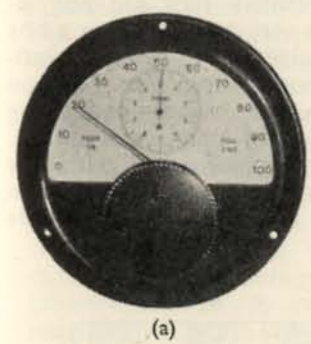


Fig. 6.9—Two types of variable capacitor bearing. Type b, in which the end plates are not under a separating stress, is to be preferred for permanence of calibration

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 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.



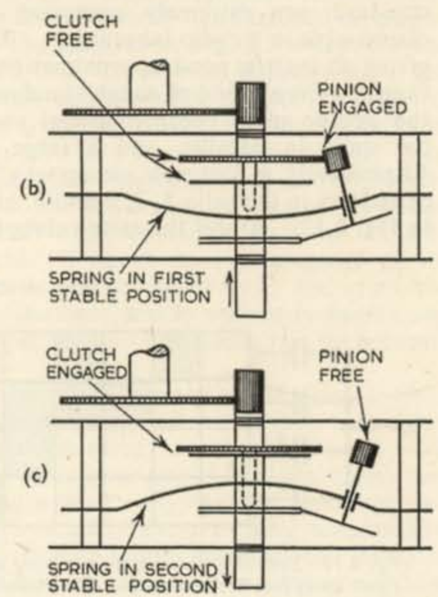


Fig. 6.10—"Microdual" two-speed instrument drive. Each division of the inner (circular) scale is one thousandth of the main scale (a). With the knob in the pulled-out position (b) there is an overall gear ratio of 120 : 1; when pushed in (c) a direct drive is substituted for one step-down, reducing the ratio to 8 : 1. (Transradio Ltd.)

Fig. 6.11—Dual-range variable air capacitor (Type C853A) with screening cover removed; maximum inaccuracy 0·03 per cent. (H. W. Sullivan Ltd.)

The Sullivan capacitors are available in multi-range patterns, even down to the least expensive grades (Fig. 6.11).

For some purposes a very small variable is required—say 1 pF. Instruments in this range are available with micrometer drive (Fig. 7.19); or an actual micrometer can be adapted to provide an extemporized capacitor, by mounting it so that its moving plunger works inside an 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. 9.32.

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 indispensable 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 decade plan, because special switches, progressively connecting the units in parallel, and a large number of units, are needed. Alternatively, a four-pole eleven-way Yaxley type of switch, with four capacitors in the ratio 1, 2, 2 and 5, gives one decade (0–10), as shown in Fig. 6.12. And if this is too elaborate, a useful unit is a single-pole

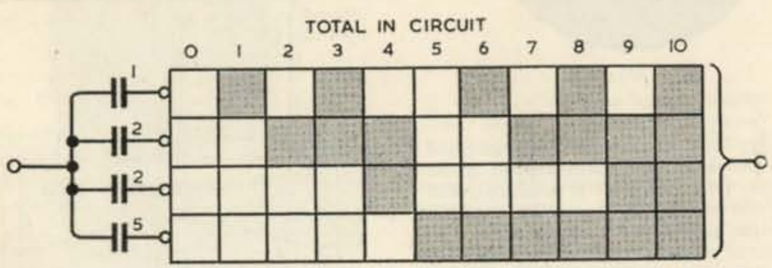


Fig. 6.12—Switching diagram for obtaining up to 10 capacitance units in steps of 1 unit, using four fixed capacitors. The shaded squares indicate switch contacts made

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.

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. 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. 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 per cent is bad; 0·2 per cent is fair; and 0·05 per cent is good. Above 0·01 μF, mica capacitors are costly, and good-quality paper types are usually substituted for ordinary purposes.

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. 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

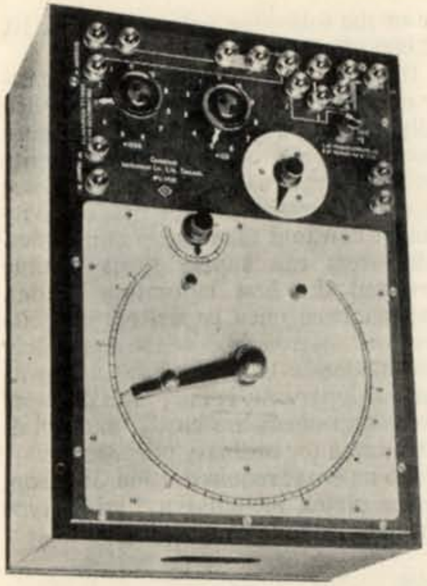


Fig. 6.13—Toroidal winding is an effective way of reducing the external field of inductors. A toroid is in effect a long solenoid bent round into a circle

Fig. 6.14—One of the types of Campbel inductometer, for use in some of he most accurate methods of inductance measurement. Note the clear, open scale. (Cambridge Instrument Co. Ltd.)

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 to wind and not very adaptable, and has a relatively high resistance (low Q). 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.

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, 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,

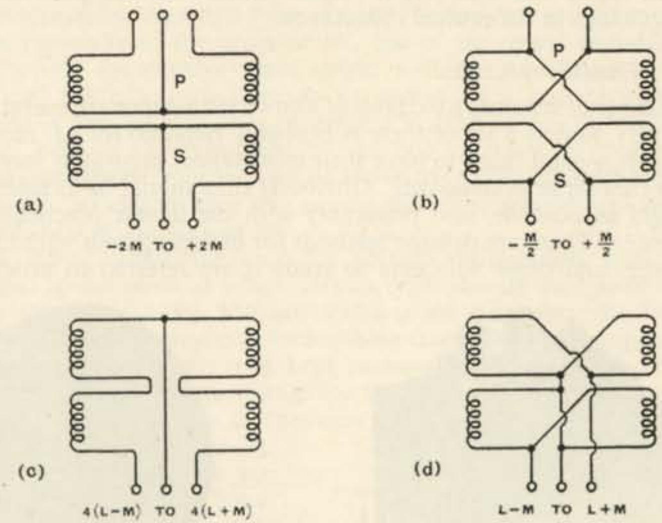


Fig. 6.15—Various methods of connecting a four-coil inductometer, giving approximately the mutual and self-inductances shown, where L is the self-inductance of one coil and M the maximum mutual between one pair of coils

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. There are definite methods for designing coils with low r.f. resistance, and those who care to study it are referred to articles by

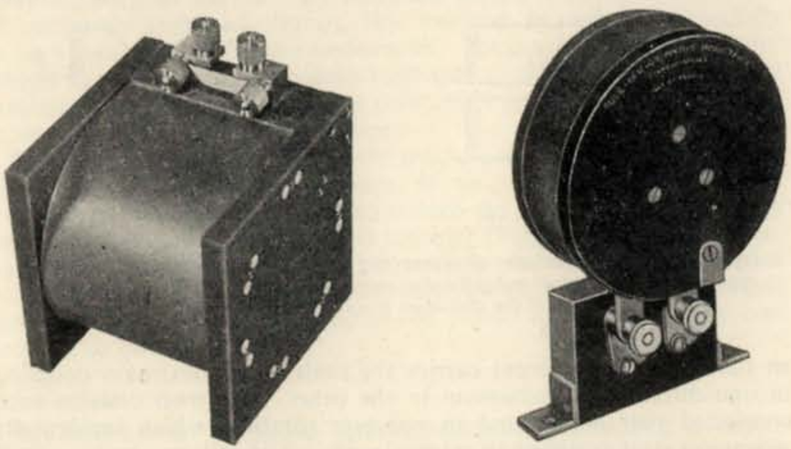


Fig. 6.16—Sullivan-Griffiths standard inductors, constructed on a compensating principle to ensure freedom from changes due to temperature. On the left is shown the first grade, for the highest precision. On the right is the second grade, giving excellent characteristics at low cost. (H. W. Sullivan Ltd.)

R. G. Medhurst in *Wireless Engineer*, February and March 1947, and a book, *The Theory and Design of Inductance Coils*, by V. G. Welsby (Macdonald, 1950). 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 and described by him in *Wireless Engineer* of October 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 per cent.

6.12. Standards of Voltage

As mentioned in Sec. 6.1, standard cells are not commonly used in radio laboratories, though if there is a call for accurate measurement of voltage, especially of the order of 1V, one of the recent portable and comparatively inexpensive types might well be the answer. Comparison of the unknown is made by means of a 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.

Neon tubes, commonly used for voltage stabilization, have been developed to the point at which certain types can be used as constant voltage standards. 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 per cent throughout life, within 0·2 per cent after the first 300 hours, and within 0·1 per cent for any period not exceeding 100 hours after the first 300 hours. The temperature coefficient is only $-$ 0·003 per cent per °C. The variation with current is about 0·35 per cent 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 this the average increase in output voltage when the input was raised from 240 V

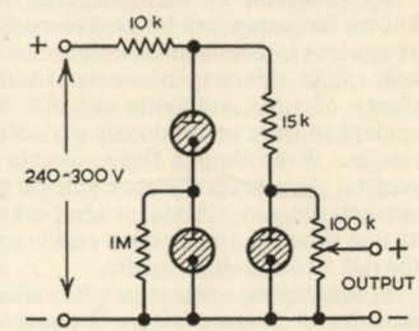


Fig. 6.17—Method of connecting 85A2 tubes to give highly stable reference voltage

to 300 V was found to be only 20 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.

Ref: "A Study of the Characteristics of Glow-discharge Voltage-regulator Tubes", by F. A. Benson. *Electronic Engineering*, September and October 1952.

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 0·000002 per cent, 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. Knowledge of frequency is now so exact as to show up 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, and it is only exceptionally that a special instrument—a wavemeter or frequency standard—has to be set apart for this purpose alone.

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 very 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 be good compared 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 10^6 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. Frequency Calibrators

The most commonly used frequency sub-standard is a plate or bar cut from a quartz crystal. It has a very sharp mechanical resonance, and as it is also piezo-electric any mechanical vibration generates corresponding differences of electrical potential between its faces, and vice versa. The equivalent electrical circuit is as Fig. 6.18, which shows

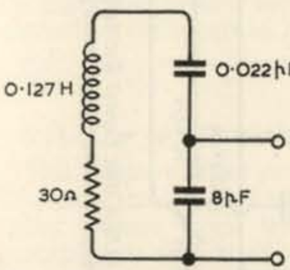


Fig. 6.18—Equivalent circuit of quartz crystal, showing typical values

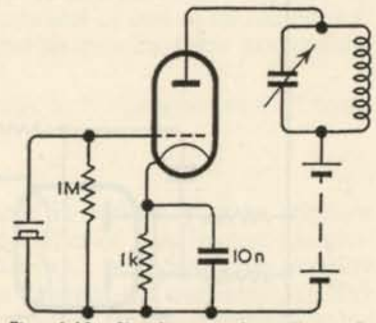


Fig. 6.19—Simple crystal oscillator for frequency checking

typical values for a crystal resonating at 3 Mc/s. When suitably connected in a valve circuit with feedback, continuous oscillations are set up at a frequency determined mainly by the dimensions of the crystal. Although crystal-controlled oscillators are used as standards of very high accuracy, it should not be assumed that any oscillator that is crystal-controlled is necessarily a highly reliable standard. Much depends on the quality and cut of the crystal; inferior specimens have a nasty trick of jumping from one frequency to another near it, and the temperature coefficient may be so high that unless the crystal is thermostatically controlled it may show little or no advantage over the best LC circuits. 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 so even without temperature control the constancy would be very good. Obviously, one would not put the crystal close to a valve or other heat-generating component.

Sometimes a crystal-controlled frequency calibrator is built into a

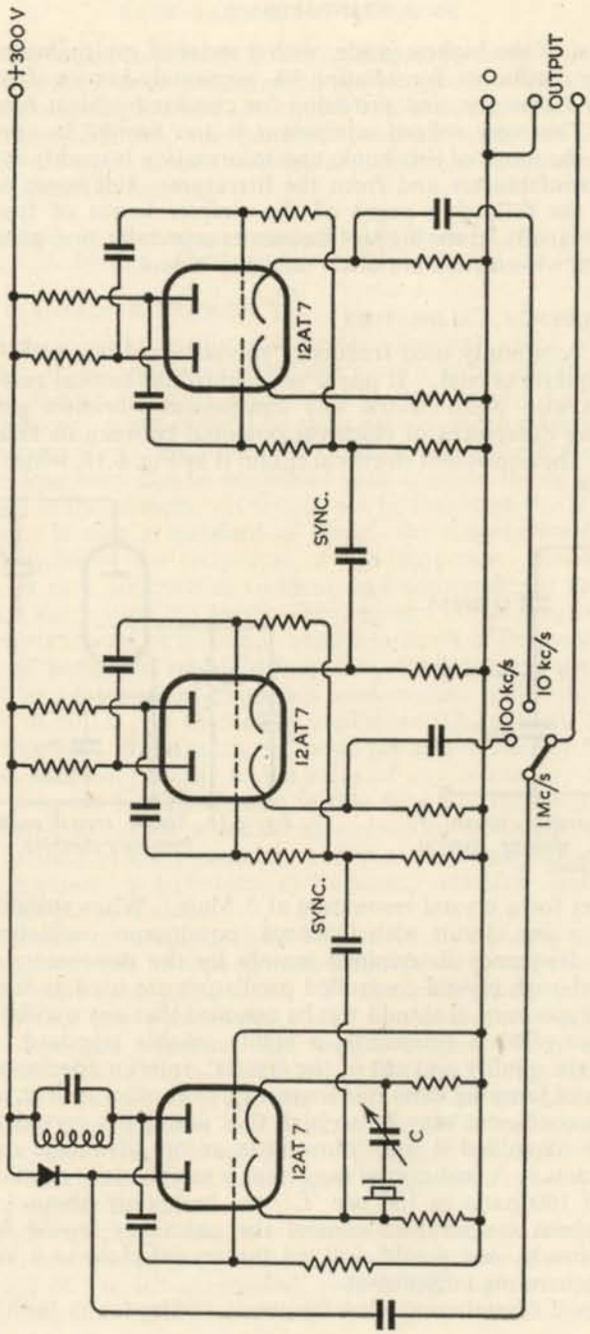


Fig. 6.20—Outline circuit of General Radio type 1213-A Unit Crystal Oscillator. The left-hand twin triode is the 1-Mc/s crystal oscillator and harmonic producer; the other two are synchronized multivibrators giving frequencies at 100-kc/s and 10-kc/s intervals

signal generator; for example, see Fig. 4.28. Or it may be a separate piece of equipment for general checking. The simplest is just a single-valve oscillator, using an adaptation of almost any of the well-known circuits. Fig. 6.19 is an elementary example. Crystals vary considerably in their " activity ", i.e., readiness to oscillate, and a circuit that suits one might not work at all with another. (It is assumed that the crystal is clean—contamination with grease seriously reduces activity—and is mounted in a holder suited to its mode of vibration.) Fig. 6.18 shows that between the only two points that are accessible the electrical coupling to the equivalent parallel resonant circuit is very loose—a low tapping on the capacitance side—and it is only because of the extremely high Q, 80,000 in this example, that oscillation is possible in an ordinary circuit. Low-activity crystals may do better if used as series resonators.* Excessive amplitude of oscillation should be prevented, because it is bad for frequency stability and may even crack the crystal.

The oscillator to be checked is adjusted to zero beat with stray pick-up from the crystal oscillator on its fundamental or one of its harmonics. To check frequencies between standard harmonics there are various methods of interpolation (Sec. 10.11).†

Ref: *Quartz Vibrators and their Applications*, by P. Vigoureux and C. F. Booth (H.M.S.O., 1950).

6.15. Use of Multivibrators

This simple nucleus can be elaborated in various ways to improve frequency stability and to extend the number of spot frequencies provided. Apart from temperature control, one frequency-stability precaution is the use of a buffer stage between the oscillator and any other controlled circuit. Such a controlled circuit is usually some device for multiplying harmonics. While a diode or other distorter can do this, a more effective means is a multivibrator (Sec. 4.31). This type of oscillator is distinguished by the vast number of harmonics it generates, and the ease with which its frequency, adjusted approximately, can be pulled into step by a suitably connected standard oscillator. Each of its harmonics is then an exact multiple of the controlling oscillator's frequency. Alternatively, it can be synchronized with every *n*th cycle of the controlling oscillator, in which case the multivibrator acts as a $n:1$ frequency divider. With a sufficient combination of both kinds, calibration points can be established at those intervals over the whole frequency range. Starting from a single 100-kc/s crystal standard, an equipment described by H. J. Finden (*Electronic Engineering*, June 1950, pp. 220–6) covers the range from 1 kc/s to 10 Mc/s in 1 kc/s steps.

* H. B. Dent, in *Wireless World*, July 1952, pp. 275–277.
" Series-resonant Crystal Oscillators ", by F. Butler. *Wireless Engineer*, June 1946, pp. 157–160.

† A practical self-interpolating crystal calibrator is described by D. Cooke in *Electronic Engineering*, January 1952.

The subject of multivibrator locking circuits is particularly well treated by F. E. Terman (*Measurements in Radio Engineering*, pp. 132–4).

Fig. 6.20 is an outline of the system used in the comparatively simple and compact General Radio crystal oscillator. The standard is a 1-Mc/s oscillator, using a coupled-cathode circuit due to F. Butler (*Wireless Engineer*, November 1944) in which the crystal acts as a series resonator. C is for slightly varying its frequency to bring it into step with a broadcast standard. As 1 Mc/s is rather a high fundamental frequency for a multivibrator (though not an entirely impracticable one), detectable harmonics up to the 1,000th are generated by a germanium diode as a distorter. The 100-kc/s and 10-kc/s multivibrators develop arrays of harmonics giving closer points up to 250 Mc/s and 25 Mc/s respectively; the appropriate array is selected by a switch.

To encourage the high r.f. harmonics, multivibrator circuits and valves should be chosen and arranged to have minimum stray capacitance.

The simpler type of multivibrator referred to in Sec. 4.31, shown here in Fig. 6.21,* has a number of advantages over the conventional sort, not the least being that its frequency is controlled by one R and C. Synchronizing signals can be fed into either grid. Full details are given by J. E. Attew (*Wireless World*, March 1952, pp. 114–116) of a decade chain of multivibrators driven from a 100-kc/s crystal, the multivibrators being a variety of this type. The 10 : 1 ratio is obviously the most convenient, but it is well known that unless special precautions are taken there is a considerable risk of confusion due to multivibrators slipping over to another ratio, say 9 : 1. The actual ratio in operation can conveniently be seen by means of the c.r. tube methods described in Sec. 10.10. In this instance the square pulses from the multivibrators are reshaped to provide a reversed pulse just before the synchronizing pulse, inhibiting any tendency to premature triggering. The 100-kc/s crystal oscillator and distorter provides its own output at that frequency, and controls separate outputs at 1 Mc/s, 10 kc/s, 1 kc/s, and if necessary 100 c/s, 10 c/s and 1 c/s.

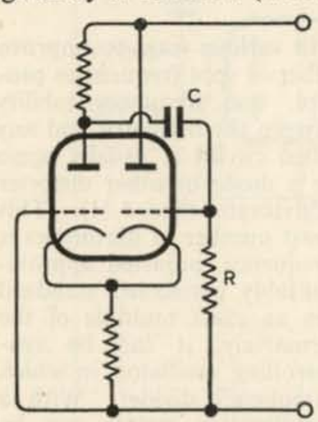


Fig. 6.21—Simple cathode-coupled multivibrator circuit

The thrifty experimenter may find that unless he is very lucky a really good crystal adjusted exactly to the frequency he wants, say 100 kc/s, is expensive. It may be easier to pick up a specimen ground

* K. A. Pullen, in *Proc. I.R.E.*, June 1946, p. 402.

to some odd—though accurately known—frequency, which can of course be used for a calibrator, but there is a sacrifice of convenience. If he is within range of the B.B.C. 200-kc/s transmitter at Droitwich, and does not insist on having standard frequency laid on during the small hours of the morning, he may be glad of the suggestion that the entire cost of a crystal can be saved by using this 200-kc/s carrier wave, suitably amplified, to control his multivibrators. In that case he has a guaranteed accuracy within 1 in 10^6, winter and summer, and a probable accuracy very much better—the error is said to be seldom more than 1 in 10^7.

6.16. Passive Frequency Standards

Although an oscillator is the most generally useful form 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 its 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 a foot or more long, with an extension control. At frequencies of hundreds of Mc/s (decimetre wavelengths) the absorption wavemeter may consist

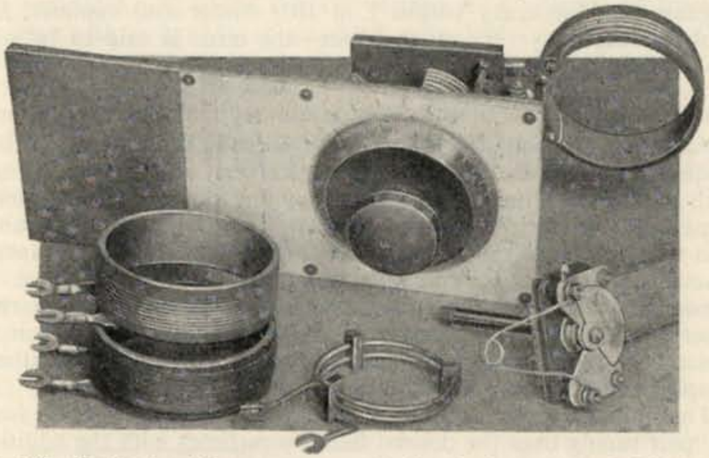


Fig. 6.22—Simple absorption wavemeters and coils; the larger covers 2–100 Mc/s and the smaller is for quick checking of decimetre waves

of a tiny capacitor with a little loop of wire connecting the terminals. Fig. 6.22 shows an example of this rigged up in a few moments, but adequate for rough comparisons; and also one covering all frequencies from 2–100 Mc/s with four coils.

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, 299,790,000) 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. The tendency, then, is to oust the term wavelength. 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.*, Pt. III, December 1945, pp. 291–5). 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

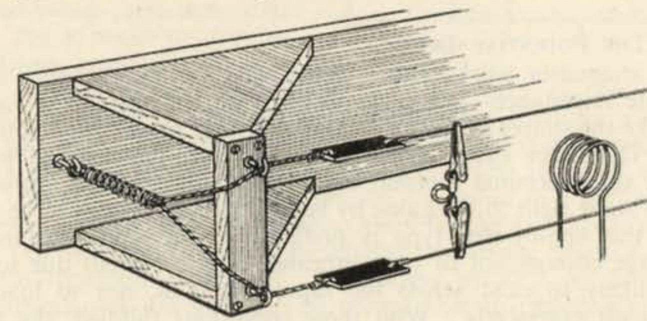


Fig. 6.23—Position of an oscillator coil for coupling to a parallel-wire resonator in wavelength measurement

position as to couple magnetically (see Fig. 6.23) 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 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 per cent.

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.

6.18. Standards of Amplification: Attenuators

There are such things as standard amplifiers, used to provide a known gain when measuring voltages too small to be read direct by valve voltmeter, cathode-ray tube, or other instrument. Their gain depends on so many things that they cannot reasonably be depended upon to high accuracy, so wherever possible a standard of loss—an attenuator—is preferred, because it consists of a passive network of resistors. If an amplifier *must* be used for measurement, its gain should be stabilized by plenty of negative feedback. While an attenuator cannot serve the purpose of an amplifier, it can be used to calibrate an amplifier (Sec. 11.6).

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 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

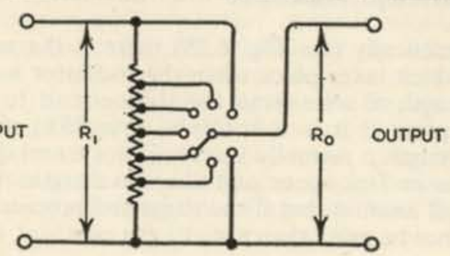
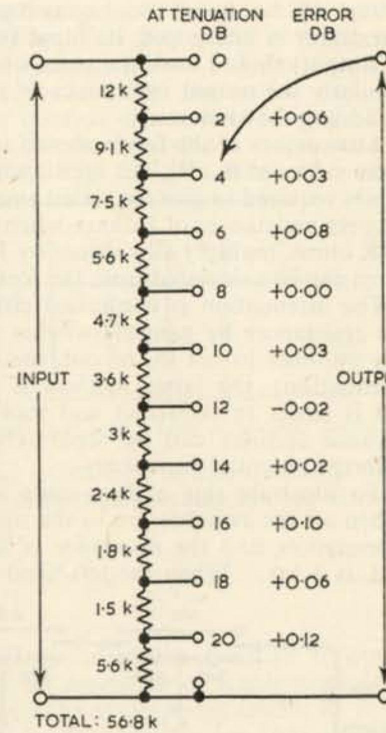


Fig. 6.24—Within certain restrictions, the simple potential divider can be used as an attenuator

Fig. 6.25—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



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_I (Fig. 6.24). 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_I. E. W. Berth-Jones has shown (*Wireless World*, February 1950, pp. 71–2) 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. 6.25 is one of his examples—a potential-divider type of attenuator totalling just over 56 kΩ and giving 0–20 db in steps of 2 db. The errors shown are in addition to those due to resistor tolerances. 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.

6.20. Matched-resistance Attenuators

The potential divider has two limitations: if the impedance into which it works is not infinite, the amount of attenuation is altered, and so is its input impedance. Wherever one part of a system is connected to another—say, aerial to tuning circuit, or amplifier to loudspeaker or telephone line—the second part presents a certain impedance (its input impedance) to the first, while the first is equivalent to a signal source working through an impedance (its output impedance) into the second. If these impedances are altered, the working of the system will be altered too, e.g. as regards frequency response. So if an attenuator is interposed, its input (with the second part connected to its output) should have the same impedance as the second part alone. Similarly the output impedance of the first part should be unaffected by adding the attenuator.

Attenuators of the forms shown in Fig. 14.36 fulfil these conditions if the values of R_1, R_2, etc. are suitably chosen. Table 14.10 shows the values required to give the stated amounts of attenuation when working between resistances of 1 ohm; when working between equal resistances of R ohms, multiply all r values by R. The values for attenuation not given can be calculated from the formulae in Fig. 14.36.

The attenuation is controlled either by simultaneously varying all the resistances by ganged switches or by using a set of change-over 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. 6.26. 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

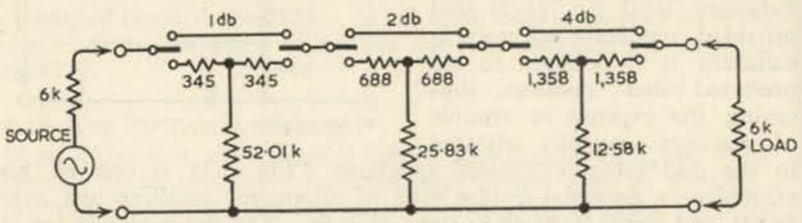


Fig. 6.26—Worked-out example of a three-stage T attenuator giving a maximum of 7 db in steps of 1 db

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. 14.36) 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 Yaxley type.

6.21. The Ladder Attenuator

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. 6.27 and 4.24) is much used, especially in signal generators and audio amplifiers. There are various slight modifications of this type: for example, the output and input can be interchanged. The ladder consists of a series of ˥ or Γ stages, which can be arranged to present a constant resistance to one end only. It is best illustrated by an example. Fig. 6.27 is a series of Γ stages,

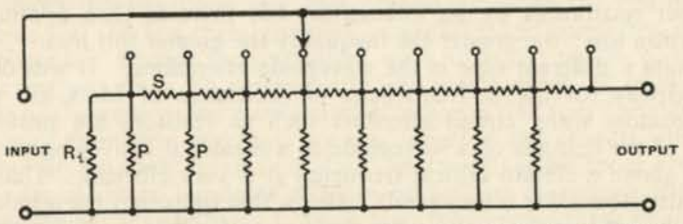


Fig. 6.27—A convenient type of attenuator much used in radio work—the "ladder"

and assuming that source and load resistances, R, are equal, and that a constant resistance is required from the *input* standpoint, the arms are calculated by:

$$\left.\begin{aligned} S &= R\,(a-1) \\ P &= R\left(\frac{a}{a-1}\right) \end{aligned}\right\} a = \frac{V_{in}}{V_{out}}$$

The resistance R_I in Fig. 6.27 is called the iterative resistance, because it is equivalent to an infinite repetition of stages towards the left. In this type of attenuator, with the assumptions stated, $R_I = aR$. For example, suppose source and load resistances are 10 Ω and each step is to be 5 db. R, then, is 10, and a is 1·778; from which S = 7·78 Ω, P = 23 Ω, and R_I = 17·78 Ω.

R_I and P in parallel are 10 Ω, which in series with S makes 17·78, equal to R_I. So R_I and the first stage together have the same resistance as R_I. This is true of any number of stages: their resistance towards the left is equal to R_I. What about the right? The load is 10, and in series with S and in parallel with P makes 10 again. So the resistance through all paths from any stud to which the source is switched (excluding the path through the source itself) is 17·78 in parallel with 10, or 6·4 Ω. This, though not equal to 10, is at least constant at all switch positions. Because it is not 10 the system is less efficient than one in which the impedances are perfectly matched, and even when the source is tapped right across the load there is a loss in the attenuator, called the insertion loss.

The attenuation per stage, working from the load end, can easily be found to be 5 db, but the output resistance is 6·4 Ω at the end stud, 12·75 at the next, 16·0 at the next, and tends towards 17·78 at an infinite number of steps away.

Nevertheless, where strict matching is not essential, it is a useful type of attenuator, and can be made quite cheaply.

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 of course the number of decibels is proportional to

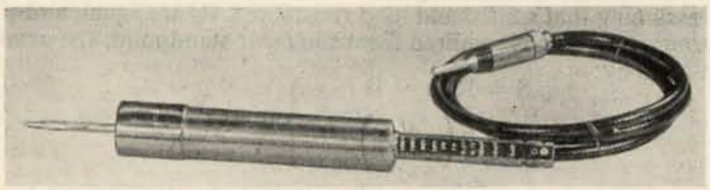


Fig. 6.28—A capacitive attenuator incorporated in the probe of a millivoltmeter, enabling the full-scale reading to be varied from 1 mV to 1 kV in 12 steps. (Philips Electrical Ltd.)

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 per cent 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.

An interesting exception is the piston attenuator used over a range of 1 kc/s to 30 Mc/s in the Philips GM.6006 millivoltmeter, where it is installed in the probe (Fig. 6.28) and serves to select any of the 12 ranges from 1 mV to 1 kV. Details are given in *Philips Technical Review*, January 1950, pp. 206–214.

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 by far the most important class. Although some types are now made for use up to hundreds of Mc/s, bridges are more generally 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 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, 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 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 100 μF as the other (a cathode-resistor by-pass). These enclose a range of a hundred thousand million to one (10^{11} : 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 Muirhead D-197-A bridge, for example, which contains everything in a box about the size of a table radio set, covers 0·001 Ω to 1 MΩ, 1 pF to 100 μF, 1 μH to 1,000 H, and dissipation factor. Standards of equal accuracy, continuously variable between those limits, would be excessively costly and occupy a large part of the laboratory.

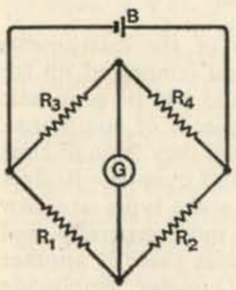


Fig. 7.1—Simple Wheatstone bridge, on which all other forms of bridge are based

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 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 $R_1 R_4 = R_2 R_3$ 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 four 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. And the same 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.

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 know which to use. The chief difficulty is deciding 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 radio 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. a 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. A.c. detectors, however, can be made so, as for example the selective amplifier-indicator described in Sec. 5.20 (1). 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. A " magic eye " without additional amplification makes a suitable detector where extreme sensitivity is not required. 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.

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

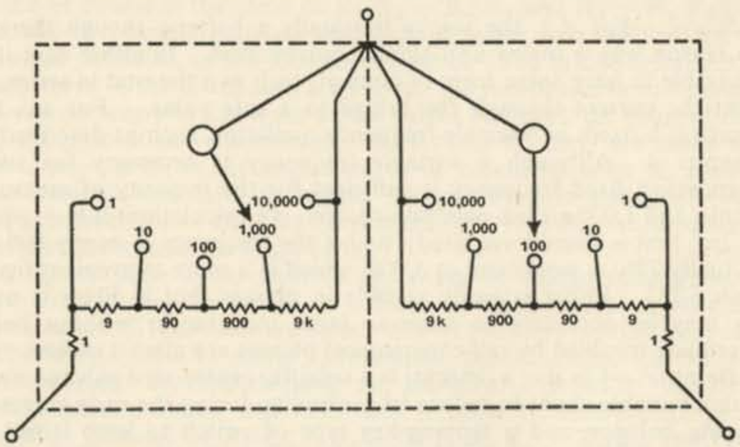


Fig. 7.2—A pair of switched ratio arms suitable as the nucleus of a general-purpose bridge

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.

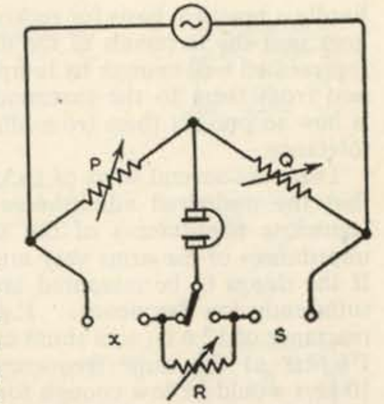


Fig. 7.3—General-purpose a.c. bridge; also suitable for d.c., if battery and galvanometer substituted for oscillator and phones. For a.c. a standard of reactance is connected at S

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.

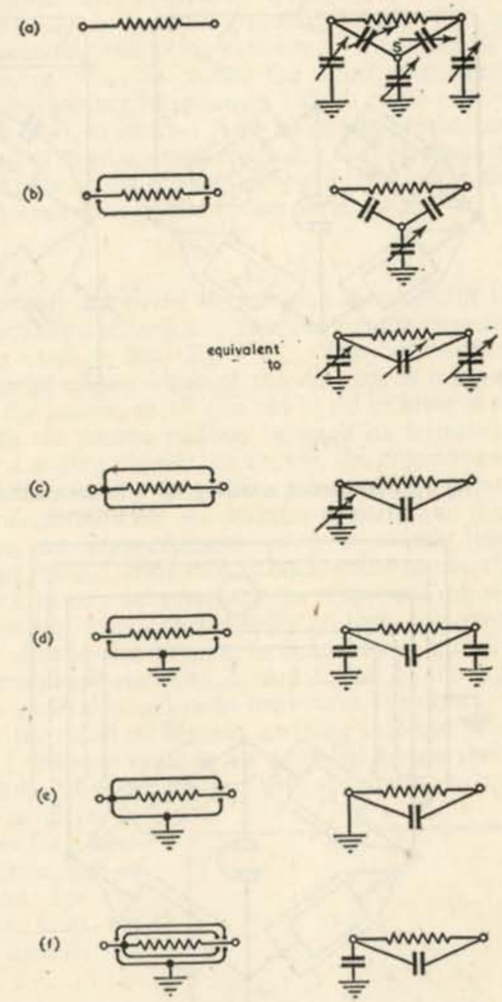
There are several ways of tackling it. One method is to make sure that the undesired admittances are all very much smaller than the legitimate admittances of the arms. In other words, to make the impedances of the arms very much smaller than those of stray paths. If the things to be measured are inductances, 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.

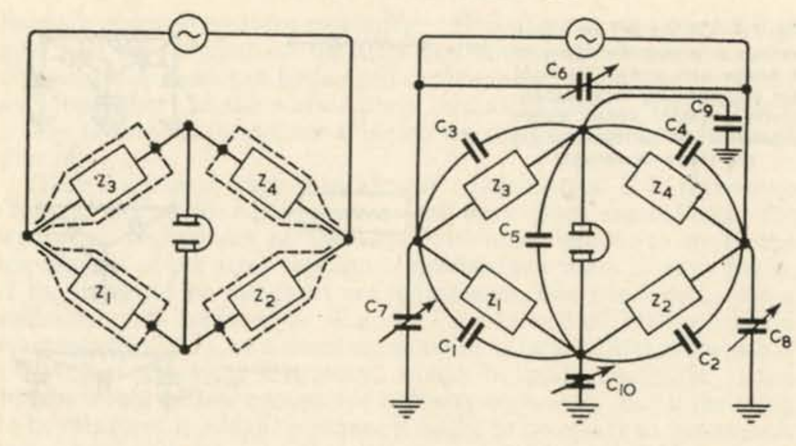
The remaining diagrams show the effects of enclosing the arm in a fixed metal screen, which then forms the only immediate surrounding. In *b*, where the screen is " floating ", its variable capacitance to earth causes a small liability to variable shunt capacitance, as shown more clearly in the second 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

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



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 pairs of screens, does not affect the readings, because it comes across the detector; nor does C_6 across the source. But the earth



(a) (b)

Fig. 7.5—(a) Partial screening of a bridge, with (b) the corresponding stray capacitances

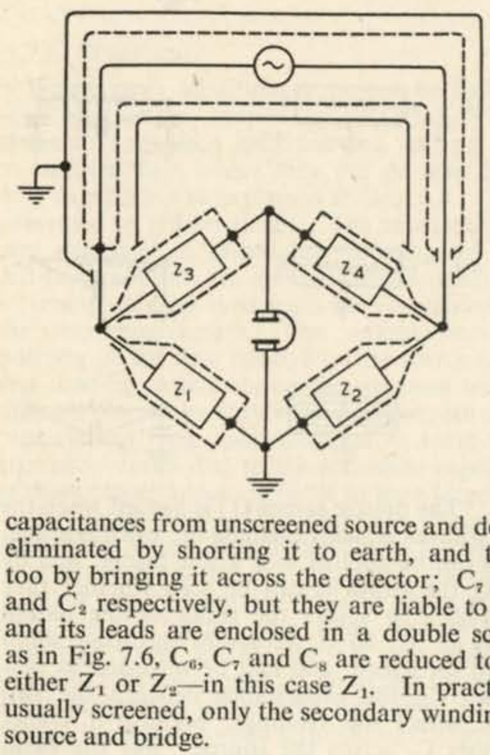


Fig. 7.6—Full screening of a bridge having one point earthed. The arm that cannot be brought to earth potential at either end (in this case the source) is screened on the Fig. 7.4f plan; the outer earthed screen is shown in unbroken line, to distinguish it from the other screens, shown dotted

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 that a 1 : 1 ratio is being used, so that $\mathbf{Z}_3 = \mathbf{Z}_4$. Then at balance, when $\mathbf{Z}_1 = \mathbf{Z}_2$, 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.

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. 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.

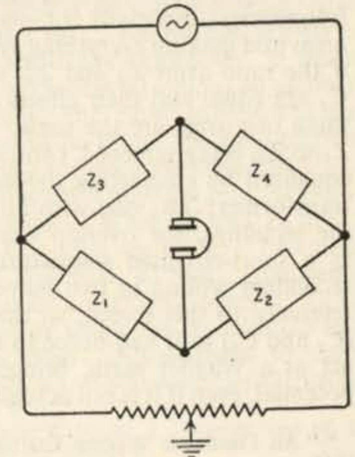


Fig. 7.7—Simplest form of Wagner circuit, by which the detector can be brought to earth potential without actual earthing, thereby eliminating the bad effects of most of the stray capacitances

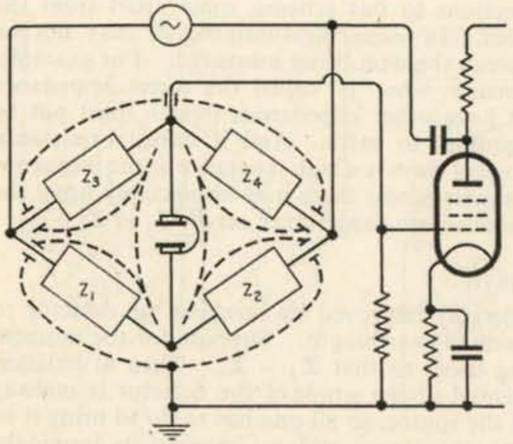


Fig. 7.8—Automatic Wagner earth system due to C. G. Mayo

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.

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, pp. 103–4.

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, 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}),

* " An Electronic Wagner Earthing Device." *Muirhead Technique*, January 1947.

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. Transformer Ratio Arms

Finally, the following method 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. The scheme is really the logical development of the one just described—a low-impedance source earthed somewhere in the middle—whereby such a source is also made to serve as ratio arms. Since the ratio of a perfect transformer is unaffected by connecting an admittance across any part of its winding (for a load on a part is equivalent to a load on the whole transformer), it is ideal for the job, and is so immune from stray admittances that one need not be bound to 1 : 1 ratio but can go to 1 : 1,000 at least. The ratio of a transformer, moreover, is determined exactly by the numbers of turns, so there is no tedious adjustment of resistance values. And once wound, the ratio cannot change. On the other hand, the transformer has to be rather specially made if it is to approximate the ideal sufficiently for accurate measurements; in particular, resistance and leakage inductance must be kept very low.

Fig. 7.9 shows how simple is the principle. When $\mathbf{Z}_1$ and $\mathbf{Z}_2$ are in the same ratio as the voltages across AE and EB, the bridge is balanced. The detector is then wholly at earth potential, so stray admittance from D to earth merely shunts it and does not affect readings. Those from A and B to earth or to one another shunt the transformer and (within reason) have no perceptible effect. Admittances to earth from either or both terminals of unknown and standard can therefore,

* " A Direct Capacitance Aircraft Altimeter ", by W. L. Watton and M. E. Pemberton. *Proc. I.E.E.*, Pt. III, May 1949, pp. 203–213.

provided they are not too enormous, be ignored. The bridge takes account only of the direct path between the two points connected.

A number of tapping-points can be provided on the transformer, to give various multiplying ratios. And, what may be less obvious, L can be measured against C, reckoned as negative L, by tapping both on the *same* side of E. 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. 7.10 shows

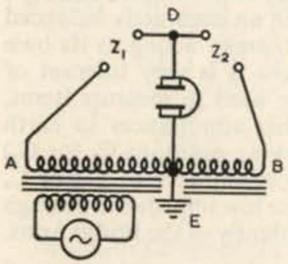


Fig. 7.9—Principle of bridge based on transformer ratio arms, which give the benefits of Wagner earthing without any adjustment

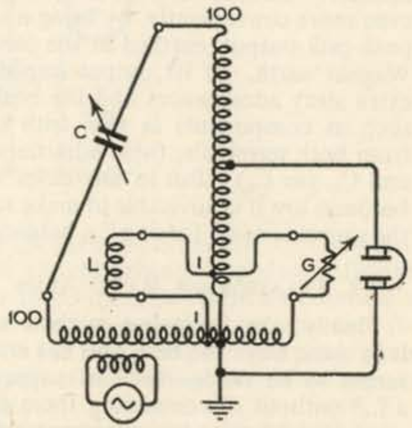


Fig. 7.10—Showing how a second transformer can be used in the detector arm to increase the overall ratio—in this example, 10,000 : 1. Inductance (L) is measured on the same side as the standard capacitance C, its resistance being balanced by the conductance G

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 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 × 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. The bridge can be earthed or not, according to requirements; generally it will need screens, connected to point E.

Both theory and practice are well expounded by H. A. M. Clark and P. B. Vanderlyn in *Proc. I.E.E.*, Pt. III, May 1949, pp. 189–202; and there is an article by R. Calvert on the application of this principle to r.f. bridges in *Electronic Engineering*, January 1948, pp. 28–29. Bridges for various frequency ranges from a.f. to v.h.f. are shown and described by H. L. Kirke in *J.I.E.E.*, Pt. III, March 1945, pp. 2–7. Commercial examples of transformer ratio-arm bridges are offered by Wayne Kerr, Cinema-Television, and E.M.I.

7.10. 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. 7.11, 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 the value of 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.

A variety of this bridge, in which balance is obtained by variable-ratio arms consisting simply of an ordinary " potentiometer ", with 50-c/s mains as source and magic eye as detector, is perhaps the most widely used type of bridge for ordinary component checking over a wide range of C and R, and constructional details, including an adaptor for measuring high-Q inductance, are given in Sec. 7.12. At such a low frequency there is little trouble with stray admittances, and little or no screening is needed.

When C is very large, *r* 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*.

* For a simple explanation of their theory by means of vector diagrams see *Wireless World*, April 1948, pp. 139–142.

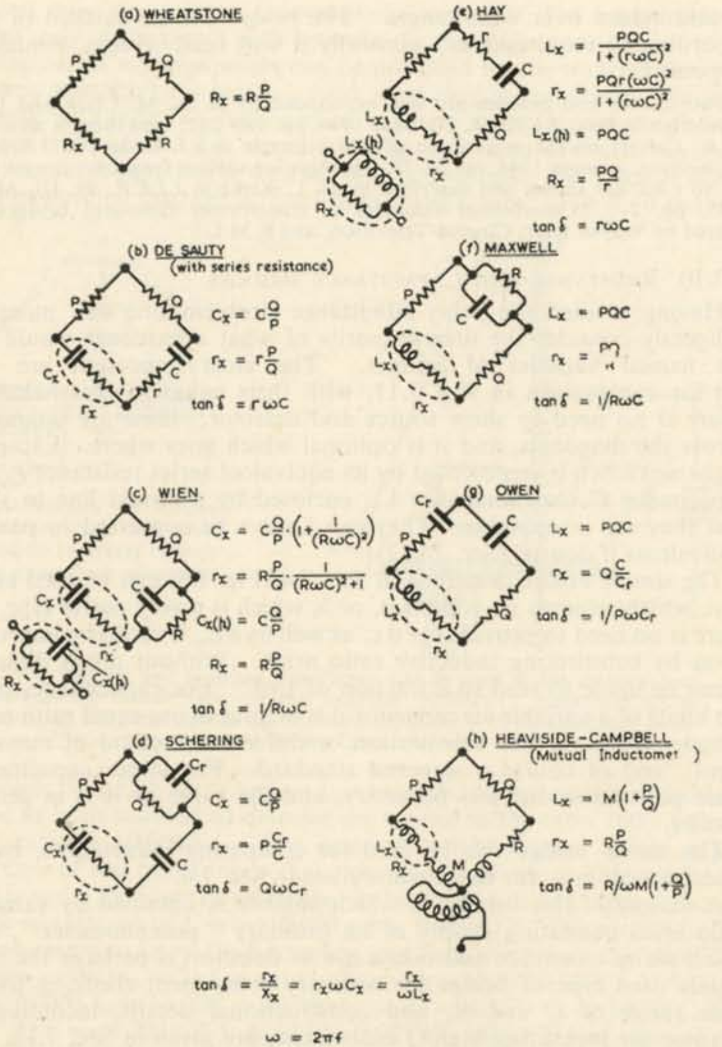


Fig. 7.11—Summary of the most-used named varieties of bridges, with their balance equations. Source and detector are omitted for clearness

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 Wien bridge is more often used with known capacitance and resistance to measure frequency than vice versa (Sec. 7.17).

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.

7.11. 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 Muirhead bridge mentioned in Sec. 7.2, covering very wide ranges of R, C, and L with sufficient accuracy for most purposes, employs types *a*, *b*, and *f* respectively, the internal rearrangements being performed by a switch. Instruments combining the same three types of bridge with the addition of *e* for high inductances, and covering similar ranges, are made by Marconi Instruments and General Radio. These wide-range laboratory bridges read also the dissipation factor or the Q of reactances.

Fig. 7.12 shows the circuit of a simple twofold bridge for measuring C and L (but not their resistance) from 50 pF to 1 μF and 10 μH to 0·1 H, with accuracy of a few per cent. Types *c* and *f* are used, and the components are lettered to agree with Fig. 7.11. A small non-inductive rheostat, r'_x, is available in series with the coil being measured in order to facilitate sharp balance. It is not used when measuring C_x. The screening case of C is joined to one side of the source, the other side of which is earthed, so the stray to earth of C comes across the source and in moderation is harmless. One side of the unknown is earthed, so if one side is screened this should be it. The capacitance of the resistors labelled R and their switching should be negligible, or at least constant, or they will upset the calibration of the standard C. A buzzer source is shown, because it lends itself to compact self-contained construction, and is cheap; but of course a valve or neon-tube oscillator has the considerable advantage of silence.

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

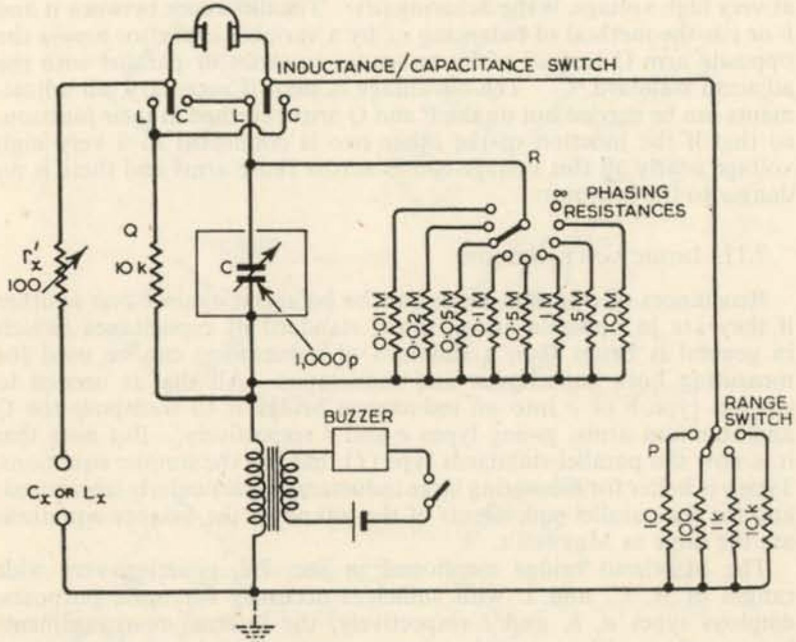


Fig. 7.12—Circuit diagram of simple portable Maxwell bridge for inductance and capacitance. The lettering corresponds to Fig. 7.11f

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.

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* (McGraw-Hill).

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. Constructional details of an improved Owen bridge are given in Sec. 7.14.

Undoubtedly 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. 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 7.11*h*) the resistance ratio arms 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.

7.12. 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 800 or 1,000 c/s, 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. 7.13 shows a circuit suitable for amateur construction, and Fig. 7.14 a complete instrument. The scale of the potentiometer, R_1 in Fig. 7.13, is calibrated to read ratios directly, as in Fig. 7.15. 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 " 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.

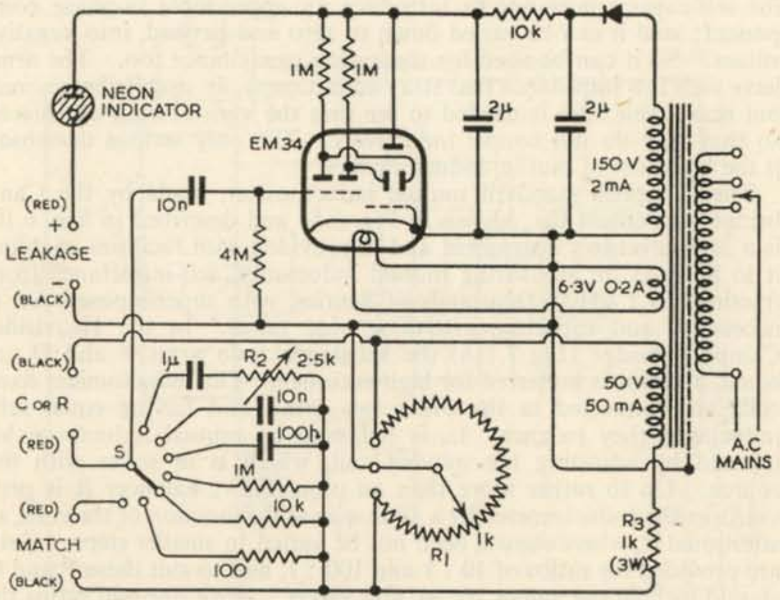


Fig. 7.13—Circuit diagram of mains-driven bridge for capacitance and resistance, using visual balance indicator. The rectifier is a Westinghouse type 16HT12

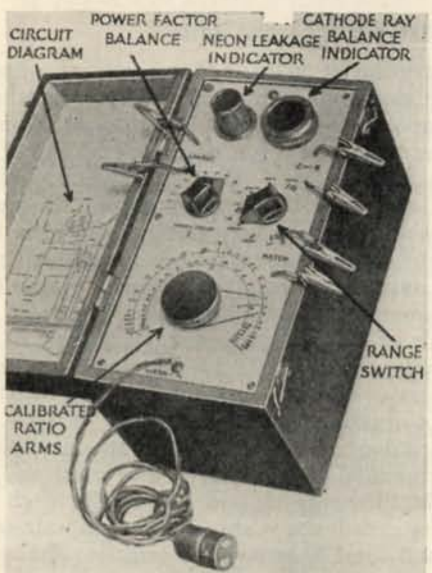


Fig. 7.14—Example of embodiment of bridge circuit shown in Fig. 7.13

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 per cent, 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.

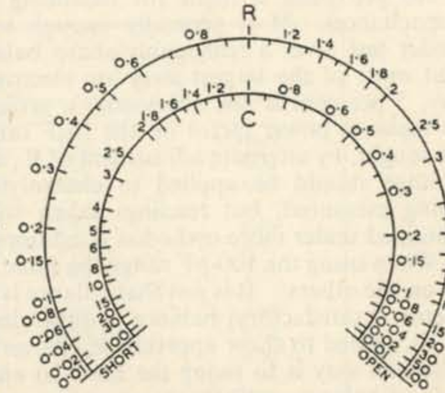


Fig. 7.15—Scale of the type of bridge in Fig. 7.13, assuming a linear R_1

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 ± 1 per cent. 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 per cent 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 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. If the fixed resistance, connected to " Match ", is 100 Ω, the other is first set to 100 Ω, giving the ratio 1 to mark on the scale. Increasing to 110 Ω gives 1·1, and so on. 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 quantity of Eureka wire can be used, on the assumption that its resistance is proportional to its length. But of course cutting the lengths of wire to find a good number of scale points is a tedious job.

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.

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

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!

An alternative design, using the neon tube as an oscillator and phones as detector, is given by H. E. Styles in *Wireless World*, March 1950, pp. 88–92.

7.13. 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 (Q large). 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

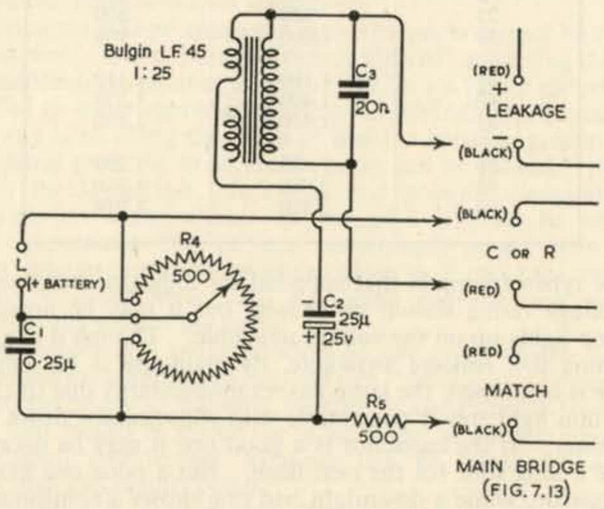


Fig. 7.16—Adaptor for measuring iron-cored inductors, showing how it is connected to the bridge shown in Fig. 7.13

as an adaptor. Assuming the latter, Fig. 7.16 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 of R_4 is not critical (nor is that of 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 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 per cent 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 $\frac{(a/b)^2}{4\pi^2 f^2 C}$ where f is 50 c/s, C is in this case 0·25 μF, and a/b is the ratio of the resistance across L to that across C and can be calibrated from the main bridge, which, of course, is itself scaled in ratios. The formula simplifies to $L = 40\,(a/b)^2$, 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.

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

Further details, and derivation of formula, are given in *Wireless World*, 12 January 1939.

7.14. 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 r.f. range. 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. In the Maxwell there is a wide disparity between the reactance to be measured and that of the standard—a calibrated variable capacitor—whereas for accuracy the arms of a bridge should be of the same order of impedance. 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. 7.11g) is shown in Fig. 7.17. 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

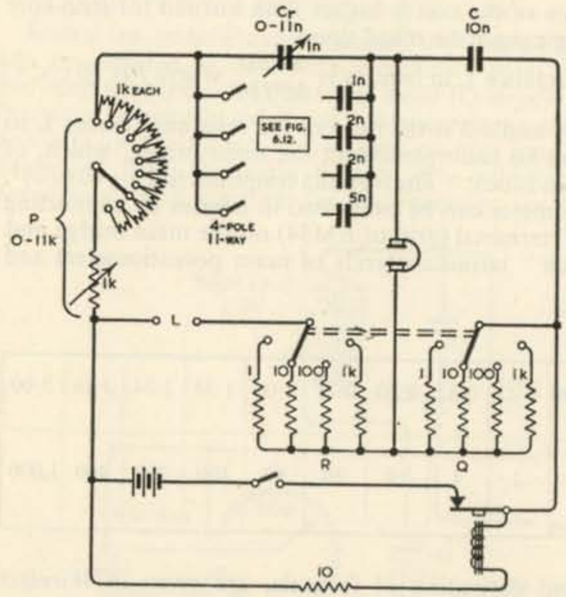


Fig. 7.17—Owen bridge for inductance, incorporating improvements by L. B. Turner. The lettering corresponds to Fig. 7.11g

may be ordinary receiver type) to total 10 nF. Fig. 6.12 shows how a four-pole 11-way Yaxley 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 on the lowest range. 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 definitely measurable.

7.15. 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 and are used in bridges manufactured by Wayne Kerr for frequencies up to 250 Mc/s. 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. *Electronic Engineering*, January 1948, pp. 28–29.
"Experimenters' R.F. Bridge", by H. V. Sims. *Wireless World*, May 1952, pp. 196–8.

Although not strictly a bridge, a type of null measuring network that is particularly suitable for high frequencies, since one terminal of both source and detector is common and can be earthed, is the parallel-T (or twin-T). The principles are explained by W. N. Tuttle in *Proc. I.R.E.*, January 1940, pp. 23–29; and details of a twin-T instrument manufactured by General Radio are given by D. B. Sinclair in the July 1940 issue of the same journal, pp. 310–8.

7.16. Influence of Source Frequency

Looking at the bridge balance conditions set out in Fig. 7.11 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. A buzzer or a neon oscillator can be used. 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. 7.11c) 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 ", *Wireless World*, January 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.

Other uses made of frequency-discriminating bridges include frequency measurement, and the filtering out of one particular frequency, as in the measurement of distortion.

7.17. Frequency Bridges and Meters

The most-used frequency bridge is the Wien (Fig. 7.11*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.

Mention should be made of direct-indicating frequency meters. Several different types are available for accurately covering a small range, such as 45–55 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 frequency is measured in terms of current and so can be read on a suitably calibrated milliammeter. One system, covering 20–10,000 c/s, was described in *Proc. I.R.E.*, April 1931; and others are produced commercially by General Radio and Airmec.

Except in special situations where continuous direct reading is required, it is generally sufficient to make use of equipment available in almost every laboratory—calibrated signal generators and oscilloscope—as described in Sec. 10.10.

7.18. 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. 14.18, and for methods of use, Sec. 9.24. The principle of most Q meters, at least for frequencies lower than v.h.f., is shown in Fig. 7.18. A known r.f. voltage is

provided by passing a standard current (monitored by the thermojunction 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

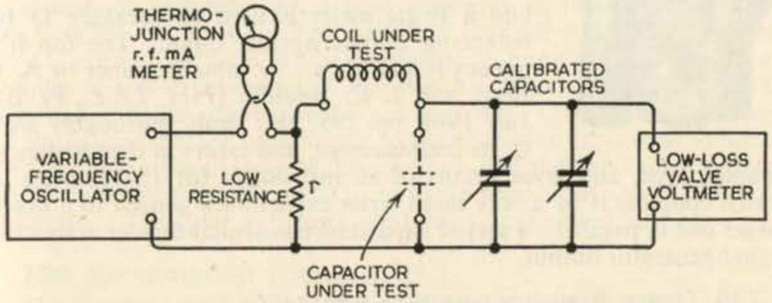


Fig. 7.18—Outline circuit of resistance-coupled Q meter. Instruments for higher frequencies use inductance in place of the low resistance r, or alternatively a capacitance-coupling system

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.

Of Q meters based on the principles described, details of the Dawe instrument are given in an informative article by H. G. M. Spratt (*Wireless World*, January 1949), and the Marconi Instruments TF.329G in a booklet *Measurements by Q Meter*, obtainable from the makers by request, which contains much general information on r.f. measurements and equipment. Among ancillary equipment is the jig shown in Fig. 7.19, which facilitates use of the Q meter for permittivity and loss tests on dielectrics. This booklet also includes the model TF.886A, one of several types that have been devised to overcome the frequency limitation imposed by the resistance *r*. It employs inductive coupling,

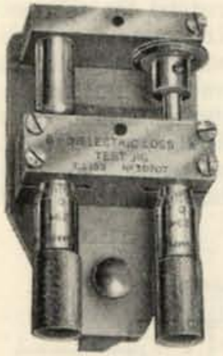


Fig. 7.19—Jig for adapting a Q meter (or other resonant measuring system) for tests on dielectric samples, which are placed between the disks on the right-hand micrometer. The left-hand micrometer controls a small increment capacitor. The system allows correction for edge effect. (Marconi Instruments Ltd.)

and a single meter is used to measure Q by balancing output against input. The top frequency is 175 Mc/s. A valuable paper by A. J. Biggs and J. E. Houldin (*Proc. I.E.E.*, Pt. III, July 1949, pp. 295–305) deals thoroughly with Q, its measurement, and errors in the resistance-coupled type, and gives details of an instrument for 15–150 Mc/s in which coupling is by a very small series capacitance ganged to a much larger one in parallel—a sort of capacitance potential-divider across the signal-generator output.

7.19. Other Resonance-method Apparatus

The types of Q meter just described are more correctly circuit-magnification meters, for they measure the voltage magnification of a resonant circuit, and only when the circuit is a theoretically simple one is this the same thing as Q. There are other methods of measuring Q, r.f. resistance, etc., for which see Secs. 9.25 to 9.30. This apparatus may generally be assembled as required, making use of an r.f. oscillator, valve voltmeter, and perhaps a specially constructed connecting unit for the resonant circuit, so it is described in the sections just referred to. A self-contained instrument working on the principle of the amount of detuning required to reduce the voltage across the resonant circuit

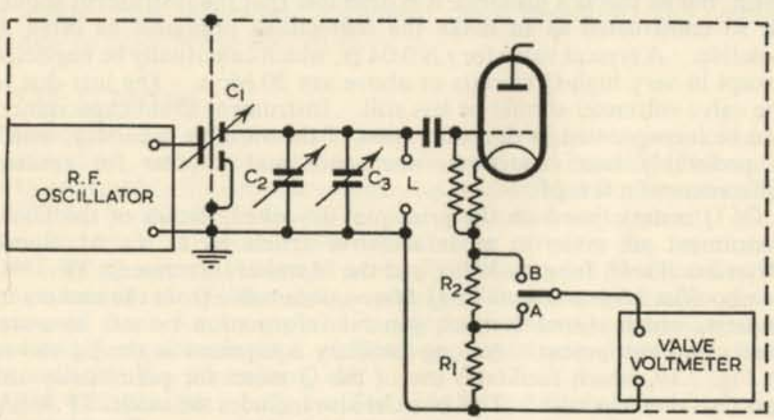


Fig. 7.20—Simple type of Q meter, due to W. R. Hinton, based on the capacitance-variation method

to 0·707 of the maximum (Sec. 9.26) has been devised, however, and lends itself to amateur construction. Details are given by W. R. Hinton in *Electronic Engineering*, October 1951, pp. 374–6, and Fig. 7.20 shows the general idea. C_1 is a special capacitor with a constant capacitance from the oscillator's point of view, but a coupling capacitance 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 ". Between the tuned circuit and the valve voltmeter is a cathode follower with a load resistance in two parts such that $R_1 = 0{\cdot}707\ (R_1 + R_2)$. The switch is first set to A, and L tuned by C_2, with C_3 at zero; and C_1 adjusted to bring the valve-voltmeter pointer to a reference mark. The switch is then moved to B, and C_3 adjusted until the original reading is restored. Q is given by the ratio of the original tuning capacitance to the change in capacitance effected by C_3. The attractiveness of this scheme is that the only things required to be calibrated are C_2 and C_3, and the ratio of R_1 to R_2 must be accurate. The valve voltmeter need not be calibrated, nor C_1.

7.20. Valve-testing Equipment

The number of valve types and the range of characteristics possessed by them are now 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. 9.34 to 9.37). 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.* And

* " Testing Steep-slope Valves ", by J. C. Finlay. *Wireless World*, March 1951, p. 108.

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.

Even a servicing tester is quite expensive if it is to be satisfactory; where funds will not run to it and some loss of convenience can be accepted, the constructional details of a simple tester for g_m, emission, and insulation given by R. E. Hartkopf in *Wireless World*, December 1946, pp. 386–390, may be of interest.

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. 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 extremely elaborate equipment, exploiting modern techniques to the full, is given by F. L. Hill and C. W. Brown of Cinema-Television in *Electronic Engineering*, November 1949, pp. 425–430. A less complicated design is described in the same journal, May 1952, pp. 220–3, by B. C. Foster; and an article on the subject by G. Bocking in *Wireless Engineer*, December 1942, pp. 556–563, includes a list of references. Still simpler apparatus that can be made up as required is shown in Sec. 9.35.

7.21. 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,

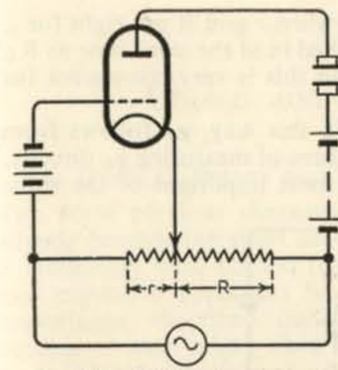


Fig. 7.21—Outline circuit showing principle of μ bridge. In practice, especially with high-μ valves, it is necessary to balance out stray capacitance, as shown in Fig. 7.22

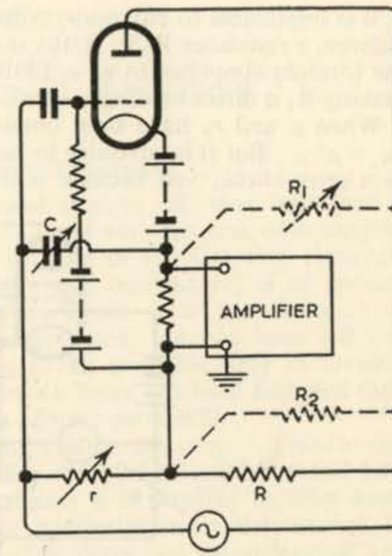


Fig. 7.22—Elaborated valve bridge, showing capacitance balance, amplifier for increasing sensitiveness of detecting balance, and additional arms for measuring r_a

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. 7.21 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. 7.22 is an elaboration of this circuit to include an adjustment (C) for balancing out reactance. A convenient value for R is 1 kΩ; with r, say 0–500 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, 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

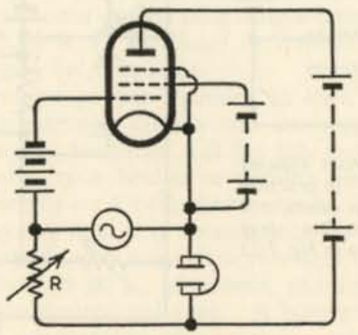


Fig. 7.23—Simple form of circuit for measuring g_m

as a guide to the condition of the valve. Fig. 7.23 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. 7.22.

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. 7.23 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 in *Electronic Engineering*, March 1952, pp. 127–9. Another system is to inject the required voltages into the most convenient parts of the valve circuit from separate transformer windings, and this—together with the General Radio valve bridge based on it—is very fully worked out and described by W. N. Tuttle in *Proc. I.R.E.*, June 1933, pp. 844–857.

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 overemphasized: " 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.

Ref: " The Activities and Equipment of an Industrial Electronics Laboratory ", by G. H. Hickling. *Electronic Engineering*, January to May 1952 (especially the May instalment).

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 4.4) to give up to, say, 100 mA at voltage variable over the range of interest (e.g., 150–450), with at least one cathode-follower tapping (Sec. 4.5) 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. 45 is recommended.

(2) *A.f. signal generator*, such as the type described in Sec. 4.21. 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. 4.23), 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. 5.21 to 5.32). Something under this heading is a necessity, even if one can only afford a second-hand tube with a home-made power unit. If it is not feasible to start off with a good one possessing all the trimmings—time base, amplifiers, etc.—room should be left for later improvements. 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 (Sec. 5.11) 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 so 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 1–20 pF, which can be calibrated from it (Sec. 9.32), covers most requirements.

(9) A *C-and-R bridge* in the class described in Sec. 7.12 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, and 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. 7.14) for inductance. R.f. coils are the chief problem, unless one has an inductometer (Sec. 7.11), 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. 7.18, 7.19 and 9.23 to 9.33). 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.

Admittedly this is not likely to suit workers handling very many valves, for whom a proper equipment (Secs. 7.20 and 7.21) 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 not yet ineffective.

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 the *Wireless Engineer* 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 an experiment* carried out by the author the standard of impedance used was 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 walnut 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.

* "Fixed Condensers for 5-metre Work." *Wireless World*, 29 September 1933.

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, some of the others definitely are. 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.

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. Thermo-couple 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 careless-ness. 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 micro-ammeter and a high voltage. Even if one takes the elementary pre-caution 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 radio 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 a small proportion of Vaseline or lanoline 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

* "Cleaning Switch Contacts", by J. J. Payne. *Wireless World*, February 1948, pp. 51–2.

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.

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.*, Pt. III, January 1950, pp. 49–55.

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 "*x*" 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. In other words, connecting the voltmeter must not lower its voltage appreciably. 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. 7.12 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, even at as low a frequency as 50 c/s, from the z.f. values.

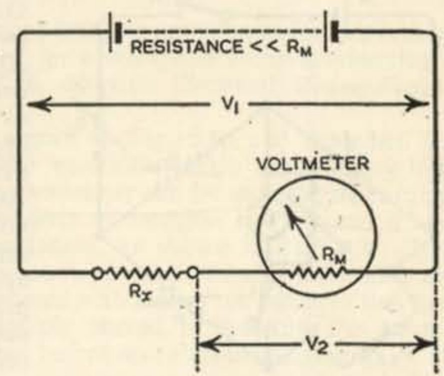


Fig. 9.1—Measurement of resistance by voltmeter

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 galvanometer 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 customary 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 $128\ \Omega$ it reads 2 to the right, one can call the resistance $127{\cdot}6\ \Omega$.

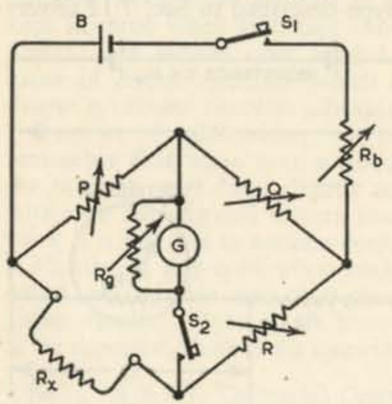


Fig. 9.2—Practical circuit of Wheatstone bridge for measuring z.f. resistance

Assuming the ratio arms P and Q are as described in Sec. 7.3, each switchable to 1, 10, 100, 1,000, and $10{,}000\ \Omega$, and R is a four-dial decade box (Sec. 6.3), variable up to $11{,}110\ \Omega$ in steps of $1\ \Omega$, one appears to have a bridge for measuring from about $0{\cdot}0001\ \Omega$ to $100\ \text{M}\Omega$. But the decade box does not provide adjustment to 1 per cent below about $0{\cdot}01\ \Omega$, 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. 5.10 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 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,

$$R_x = R_m \left(\frac{I_2}{I_1 - I_2} \right)$$

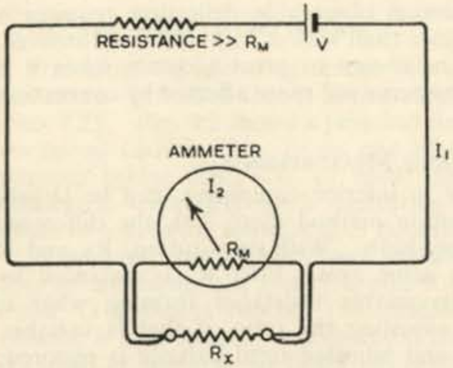


Fig. 9.3—Measurement of low resistance by shunted-meter method. Note arrangement of connecting leads to avoid displacement of zero

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 25–2,500 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. 5.15. 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 in *Electronic Engineering*, November 1952, pp. 500–3.

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 a bridge described by A. B. Boff (*Electronic Engineering*, July 1950, pp. 286–290) covers 1 to 100 million megohms. It uses a special electrometer valve voltmeter as detector, and of course every precaution against leakage.

An important thing to remember in high-resistance measurement is that most insulators and semi-conductors 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. High resistance also depends very much on temperature and humidity.

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. 5.13 and 5.14), 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. 5.15) as in Fig. 9.4. The anode voltage should be kept down to about 50 V, 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, 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.

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

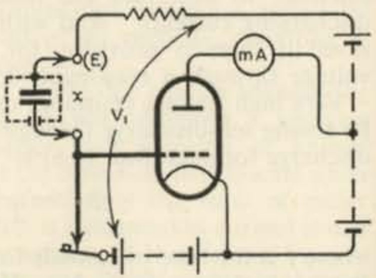


Fig. 9.4—Valve circuit for measuring capacitor leakage and other high resistance

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,

$$R_x = \frac{R_1 R_2}{R_2 - R_1}$$

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.

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_I 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 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.

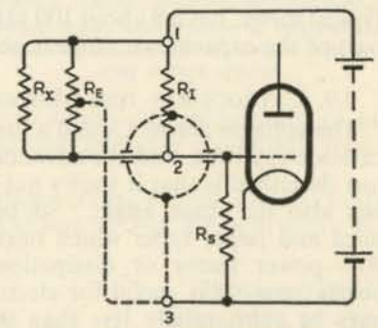


Fig. 9.5—The dotted lines show guard arrangements to intercept leakage by paths it is not desired to measure. R_I represents internal leakage and R_E external leakage

For a fuller exposition, see " Measuring High Resistance ", *Wireless World*, June 1952, pp. 236–8. A good example of guard-ring technique is the English Electric insulation tester mentioned in Sec. 9.5.

9.8. 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 $$C_x = \frac{159\ V}{fR_m\sqrt{V_1^2 - V_2^2}} \qquad [c/s;\ k\Omega;\ \mu F]$$

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 (2–100 nF).

9.9. 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 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

$$C_x = 1/\omega R$$
$$= 159/fR \qquad [c/s;\ k\Omega;\ \mu F]$$

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

Table 9.2

C_x (μF)	R (kΩ)
0·1	31·8
1	3·18
8	0·398
25	0·127
100	0·0318

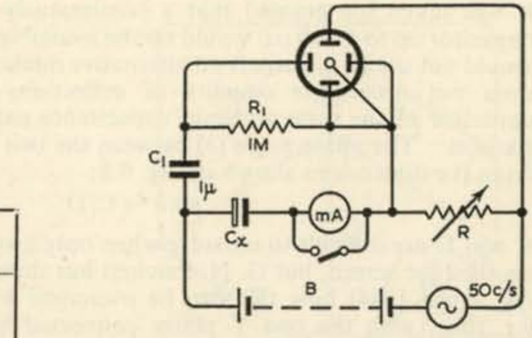


Fig. 9.6—Method of measuring capacitance, phase angle, and leakage of electrolytic capacitors by cathode-ray tube, with polarizing voltage applied

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.

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

$$\tan\frac{\phi}{2} = \frac{A}{B}$$ (*Wireless World*, 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

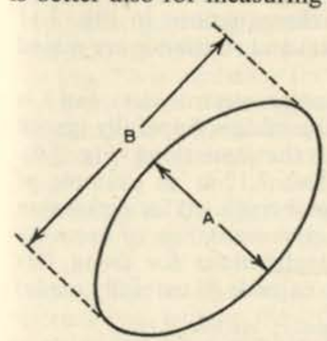


Fig. 9.7—Dimensions of c.r.t. ellipse from which phase angle can be calculated

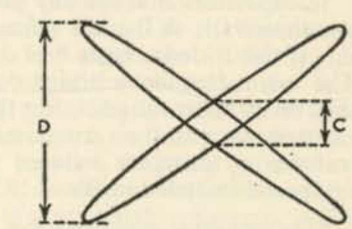


Fig. 9.8—Double-beam c.r.t. trace, more convenient than Fig. 9.7 for measuring

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 (*Electronic Engineering*, 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, 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. 9.17.

9.10. Bridge Methods for Capacitance

The three most important basic types of bridge for measuring capacitance are shown in Fig. 7.11: *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. 7.9). (In addition type *h*, being reversible, can be used to measure capacitance as negative inductance; see end of Sec. 9.20.) 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 expression, and this remains as shown in the equations in Fig. 7.11 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. 7.12 is an example of 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. 7.12. Although capable (if carefully made)

* For measuring capacitance, but not for frequency; see Sec. 7.17.

† 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° − φ, where φ is the phase angle between voltage and current.

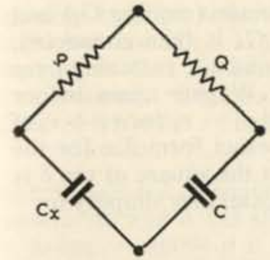


Fig. 9.9—If capacitor resistance can be neglected, the three bridges in Fig. 7.11 b–d all reduce to this

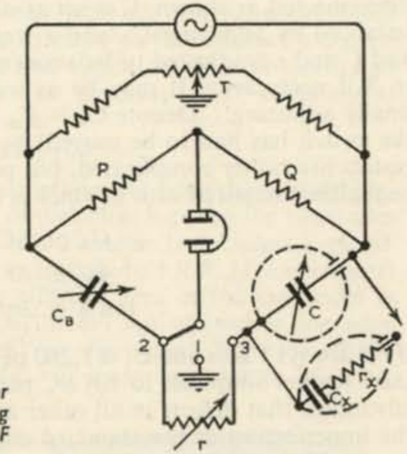


Fig. 9.10—Arrangement of bridge for difference-substitution method of measuring capacitance and loss angle up to about 1,000 pF

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 tanδ is of greater interest even than the capacitance.

9.11. The pF Range

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 $\Delta r/\Delta C$ per cent. 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.* C_x ought to be in position as shown in Fig. 9.10 while the first reading is being taken: connected at the earth end, 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.

If a permanent set-up is required for precise measurements, with emphasis on tan δ, the Schering bridge (Fig. 7.11*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.

9.12. Testing Dielectric Materials

The methods just described are suitable for measuring the loss tangent (tan δ) and dielectric constant or specific permittivity (κ_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

* "Connection Errors in Capacitance Measurements", by R. F. Field. *General Radio Experimenter*, May 1947.

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 κ_s are the true properties of the material. Detailed procedure for ensuring high accuracy is given in B.S. 903. The formula for κ_s is

$$\kappa_s = 11{\cdot}3\ C_x\, t/A$$

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. Unless a guard-ring device (described in books such as Hartshorn's *R.F. Measurements*) is used to eliminate edge effect, the effective area of the electrodes is appreciably greater than their actual area. An allowance for edge effect can be made by reckoning D as the actual diameter of the electrodes, d, plus t/π, as in Fig. 9.11. If many tests are to be done,

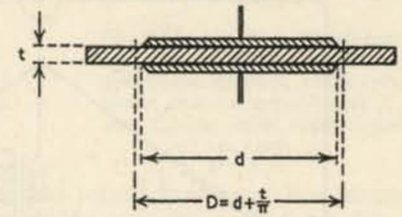


Fig. 9.11—Edge effect in tests on samples of dielectric material can be roughly allowed for by taking D as the diameter rather than the actual diameter d

calculation is eased by making A=113 sq cm so that the formula reduces to $\kappa_s = C_x\, t/10$.

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 much easier to measure at r.f. For r.f. methods see Secs. 9.24 to 9.32.

9.13. Larger Capacitances

As a continuously variable capacitance standard is not usually available much above 1 nF, the 1:1 ratio bridge methods in Sec. 9.11 may not be practicable for the higher ranges, and it is necessary either to use the variable air standard with higher ratios (e.g., the bridge in Fig. 7.12) or a fixed standard with continuously variable ratio (e.g., Fig. 7.13). 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

$$C_x = \frac{C_1 C_2}{C_1 - C_2}$$

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.

9.14. Electrolytic Capacitors

Fig. 9.12 shows modifications for applying a polarizing voltage to an electrolytic capacitor while it is being measured. Incidentally the

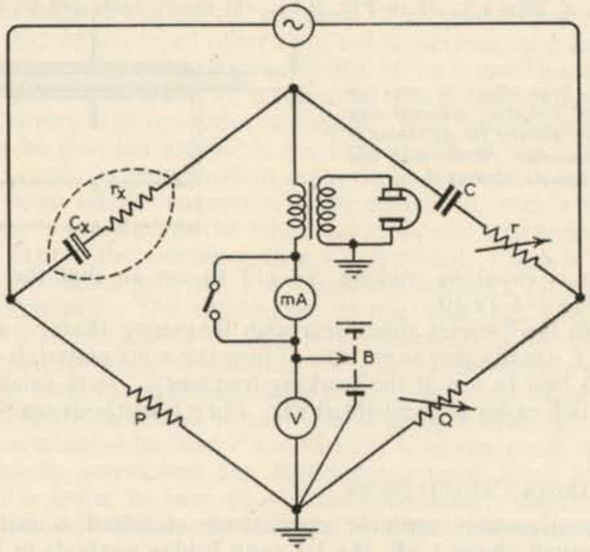


Fig. 9.12—Modifications of capacitance bridge to allow a known polarizing voltage to be applied to the capacitor under test

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 short-circuited during charging. In other respects operation is the same as for the Sec. 7.12 bridge. See also Sec. 9.9.

9.15. 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. 9.13. 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

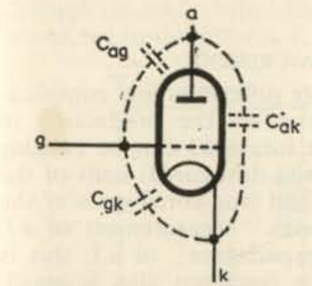


Fig. 9.13—Nomenclature of the three interelectrode capacitances of a triode

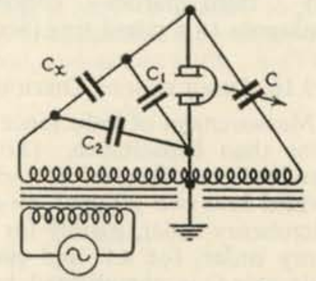


Fig. 9.14—Use of a transformer ratio-arm bridge for measuring direct capacitance (C_x) of a component that has capacitances (C_1 and C_2) or other admittances to earth

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:

$$C_1 = c_{gk} + c_{ak}$$

$$\text{Similarly} \quad C_2 = c_{ag} + c_{ak}$$

$$C_3 = c_{gk} + c_{ag}$$

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. 7.9.). In Fig. 9.14, 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.*, Pt. III, May 1949, p. 189.

If, as often happens, one terminal of C_x is earthed, the earth shown in Fig. 9.14 should be moved to the other end of the detector, leaving any "third parties" connected as shown. This connection is analogous to a guard ring (Sec. 9.7).

9.16. 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-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.

9.17. 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. 9.8), in which a 90° phase angle could reasonably be assumed.

The c.r. tube method (Sec. 9.9) 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. 9.15. 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

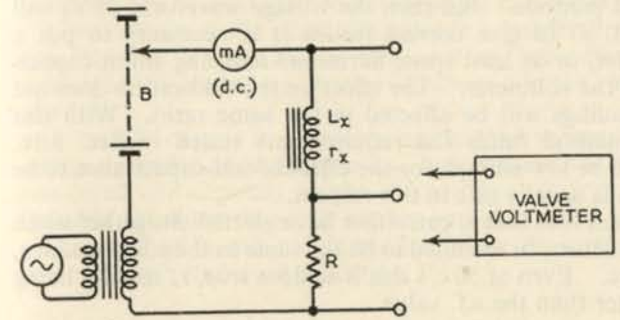


Fig. 9.15—The three-voltage method of measuring reactance and phase angle, shown here applied to an iron-cored inductor with superimposed d.c.

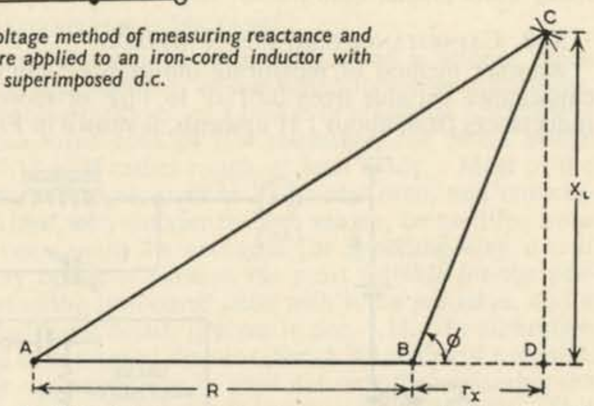


Fig. 9.16—Diagram used for deriving reactance (X_L) and phase (ϕ) from voltmeter readings in Fig. 9.15

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. 9.16 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 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. 9.16. 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.

9.18. Capacitance-comparison Method

Another method of measuring inductance,* 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. 9.17, again with

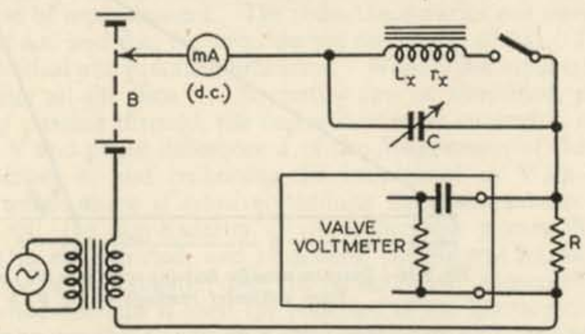


Fig. 9.17—H. M. Turner's method of measuring inductance, using large variable capacitance

provision for 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

$$L_x = 1/8\pi^2 f^2 C \qquad \text{[C in farads]}$$

At 50 c/s this simplifies to

$$L_x \simeq 5/C \qquad [\text{C in } \mu\text{F}]$$

* H. M. Turner, *Proc. I.R.E.*, November 1928, pp. 1559–69.

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. 9.17) 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.

9.19. Bridge Methods for Inductance

For reasons already mentioned (Sec. 9.16) 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. 7.13 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. 7.11. In all bridges used for iron-cored coils a tuned detector (Sec. 5.20) 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 in *Electronic Engineering*, October 1951, N. H. Crowhurst deals with the analysis of iron-core losses using a Maxwell bridge and oscilloscope.

Coming now to air-cored (including iron-dust-cored) coils, usually quite a high standard of accuracy is wanted, because they are used for tuning or filtering specific frequencies. This standard is hardly obtainable with the simple form of Maxwell bridge described in Sec. 7.11, which however is useful where the tolerance can be several per cent. The modified Owen bridge of Sec. 7.14 is capable of better results, without being unduly expensive. For high accuracy there is probably nothing to equal a bridge built around the standard mutual inductometer (Secs. 6.10 and 7.11).

The formulae for these various bridges are given in Fig. 7.11, *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 an 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.

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 dust cores and the cans are well spaced from them, 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 per cent higher than the true L_x. At one-twentieth f_r, the correction is 0·25 per cent, 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 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. 7.11*f*) in which

P and/or Q are varied. If R and C are varied, they balance r_x and L_x independently and the difficulty does not arise.

As Fig. 7.11 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. 7.11*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 per cent; 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.

9.20. High Inductance by Dye's Shunt Method

This is illustrated in connection with a Heaviside-Campbell inductometer bridge (Fig. 9.18), to which, since it is a very accurate low-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

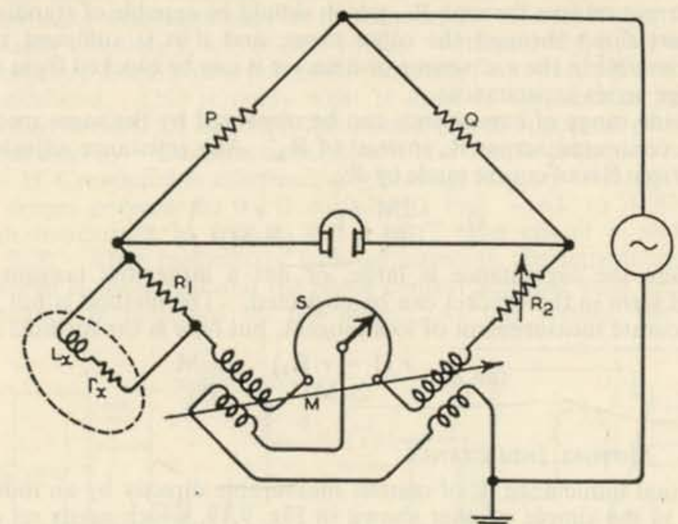


Fig. 9.18—Dye's method of using inductometer or other bridge to measure inductances much higher than its calibrated range. The same method can be used for capacitances

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 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.

$$C_x = \frac{2M}{R_2^2}\left(1 + \frac{r^2}{(2\omega M)^2}\right)$$

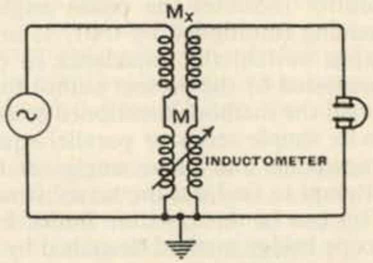
Unless the capacitance is large, or has a large loss tangent, the second term in the bracket can be neglected. The method is not ideal for accurate measurement of loss tangent, but here is the formula:

$$\tan\delta = \frac{r\,(1 - r/R_2)}{2\omega M} - \frac{2\omega M}{R_2}$$

9.21. Mutual Inductance

Mutual inductance is, of course, measurable directly by an inductometer in the simple manner shown in Fig. 9.19, 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

Fig. 9.19—Measurement of mutual inductance by direct balancing against mutual inductometer



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.

9.22. 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. 9.16, 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. 9.9 and 9.17, but some kinds of apparatus give Z and ϕ more directly. Constructional details of a simple example are given by N. H. Crowhurst in *Electronic Engineering*, January 1949, pp. 22–24. 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. 9.20. 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

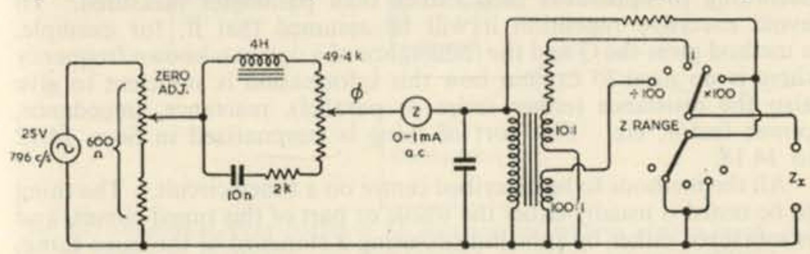


Fig. 9.20—Crowhurst's direct-reading impedance and phase-angle meter

control indicates the phase angle of the 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.

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 described by Thomas Roddam in *Wireless World*, January 1950. 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*

9.23. 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. 7.15) 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 14.18.

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. 9.21):

(*a*) In the Q-meter type, the distinguishing feature is the definiteness of the coupling from signal source to tuned circuit, whereby the amount 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

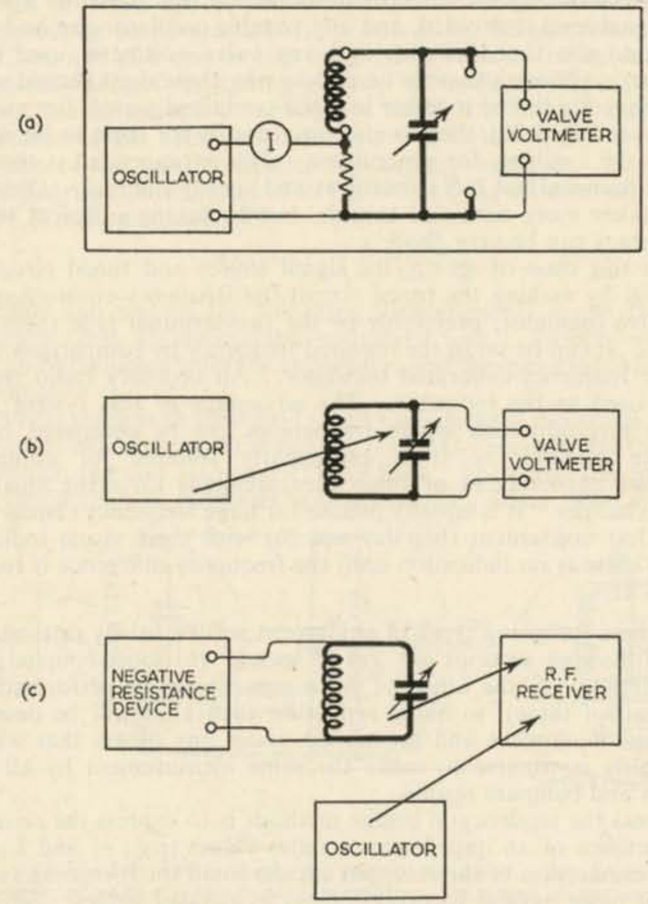


Fig. 9.21—Outline of three classes of r.f. measurement methods using a tuned circuit, shown in heavy line: (a) direct-reading Q meter; (b) loose-coupled oscillator; and (c) dynatron

marked to read Q directly, and the whole apparatus can be made up as a self-contained instrument (Sec. 7.18), 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 (an example is shown in Fig. 7.20), 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. 4.13 to 4.15). 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.

9.24. Measurements by Q Meter

Although the type diagram (Fig. 9.21*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 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. 9.22*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 (which $=\omega L_x/Q_x$), 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. 9.23*a*). Then find the frequency f_0 at which the coil resonates with its own C_0 alone; that is to say,

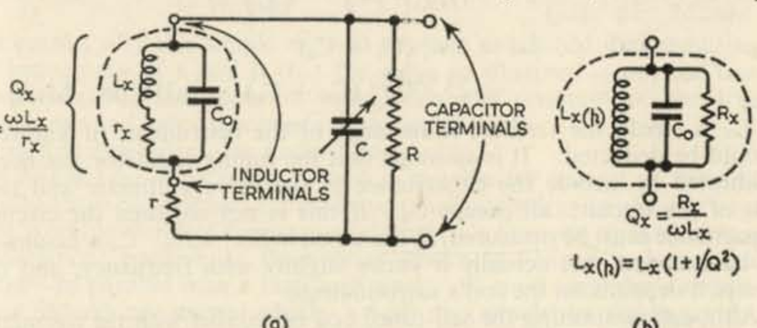


Fig. 9.22—Symbols used to denote the parts of the tuned circuit in place in a Q meter, of which r is the coupling impedance. The dotted circles enclose the alternative equivalents—series and parallel—of the inductor being measured; C is the calibrated tuning capacitor; and R the equivalent parallel resistance of the voltmeter, etc.

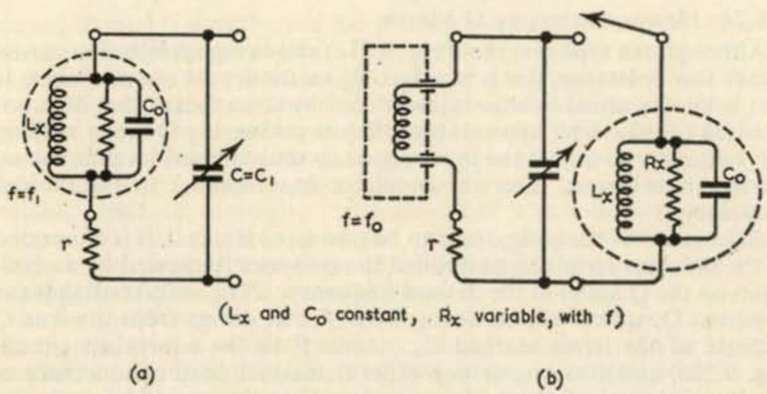


Fig. 9.23—Two stages in measuring self-capacitance (C_0) and inductance (L_x) by Q meter

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 terminals) as in Fig. 9.23*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. 9.23*b*). This frequency is f_0.

$$\text{Then} \qquad C_0 = \frac{C_1}{(f_0/f_1)^2 - 1} \simeq \left(\frac{f_1}{f_0}\right)^2 C_1$$

$$\text{and} \qquad L_x = 1/\omega_1^2\,(C_1 + C_0)$$

$$= 25{,}330/f_1^2\,(C_1 + C_0) \qquad [\mu\text{H};\ \text{pF};\ \text{Mc/s}]$$

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. 9.26. C_0 is assumed to be constant, but actually it varies slightly with frequency, and of course it depends on the coil's surroundings.

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

$$R_x = \frac{Q_1Q_2}{2\pi f_0 C_1(Q_1 - Q_2)} \qquad [\text{M}\Omega;\ \text{pF};\ \text{Mc/s}]$$

$$\text{and} \qquad Q_x = \frac{C_0Q_1Q_2}{C_1(Q_1 - Q_2)}$$

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. 9.22*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. 9.23*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 the order of 0·02–2 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

$$25{,}330/f^2(C_2 - C_1) \qquad [\mu\text{H};\ \text{pF};\ \text{Mc/s}]$$

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

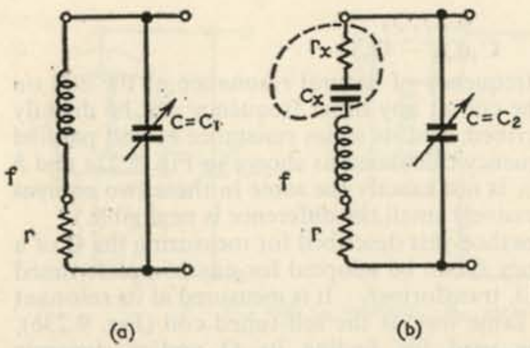


Fig. 9.24—Two stages in the series method of measuring low impedance, here typified by a large capacitance C_x with its losses represented by r_x

frequency—close enough not to let any crevasses go unnoticed—and the results plotted. Some examples are given in *Wireless World*, November 1951, pp. 494–8.

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 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:

$$R_x = \frac{Q_1Q_2}{2\pi f C_1(Q_1-Q_2)} \qquad Q_x = \frac{C_xQ_1Q_2}{C_1(Q_1-Q_2)} \qquad \tan\delta_x = \frac{1}{Q_x}$$

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. 9.24). 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

$$L_x = (C_1-C_2)/\omega^2C_1C_2$$

Dielectric Materials.—A sample is prepared as described in Sec. 9.12, 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. 7.18), which includes details of a special micrometer test jig (Fig. 7.19) and edge-effect corrections. The sample is measured as a capacitor by the substitution method, and the formulae already given apply:

$$C_x = C_1 - C_2 \qquad \tan\delta_x = \frac{(Q_1-Q_2)C_1}{Q_1Q_2C_x}$$

The specific permittivity or dielectric constant is

$$\kappa_s = 11{\cdot}3\, C_x\, t/A \qquad [\text{pF; cm}]$$

where t is the thickness and A the effective area of the sample between the electrodes.

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. 9.29 are preferable, because zero beat note can be located much more precisely than the exact peak of resonance.

9.25. "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 calibrations are required, and as they are quite independent 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 an entirely reliable professionally-made Q meter, the scheme indicated in Fig. 9.21*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. 12.9.) 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. A good example of a self-contained instrument in this class, by W. R. Hinton, is referred to in Sec. 7.18, and is easier and cheaper to make than the commercial types of Q meter.

Simple though the 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.

9.26. 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). 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. 9.25. 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

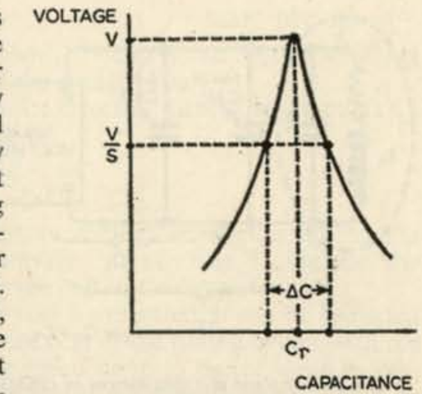


Fig. 9.25—Showing the values measured in the capacitance-variation method of measuring Q, etc.

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 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; Hinton uses 0·02–0·5 pF (Sec. 7.18).

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. 9.32.

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

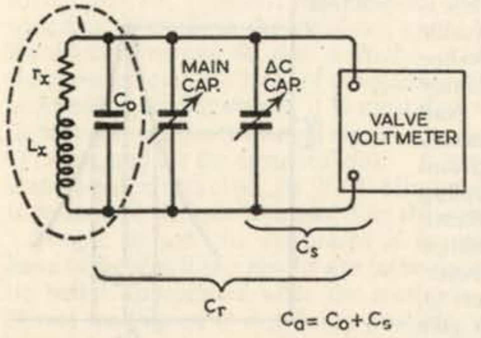


Fig. 9.26—Analysis of capacitances in capacitance-variation method of measurement

coil, C_0. What one wants to know is how much to add to the main capacitor reading to give C_r. Let us call it C_a (Fig. 9.26). 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 per cent 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-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. 9.27. 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,

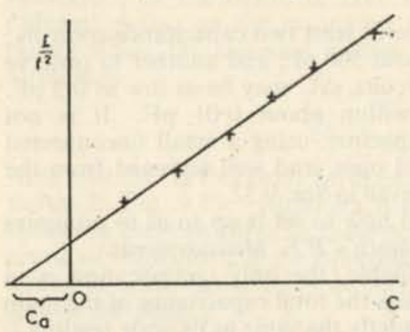


Fig. 9.27—Derivation of coil and circuit capacitance (C_a) from readings of resonant frequencies at a number of values of added capacitance C

$$L_x = 25{,}330/f^2 (C + C_a) \qquad [\mu H;\ pF;\ Mc/s]$$

If the graph Fig. 9.27 has been drawn, $1/f^2$ and C can best be obtained from any point near the top end of the straight line.

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 per cent 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. 9.24, 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

* See " A Method of Measuring the Self-capacitance of Coils ", by M. G. Scroggie. *Wireless Engineer*, September 1933, pp. 477–9.

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 $V_1\Delta C_1 = V_2\Delta C_2$.

9.27. 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 oscillator varied to find a point each side of the peak where the voltage across the circuit is divided by S (Fig. 9.28) 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 unless Q_x is very low indeed—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_r. A number of frequencies are known to 1 in 10^6 (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.

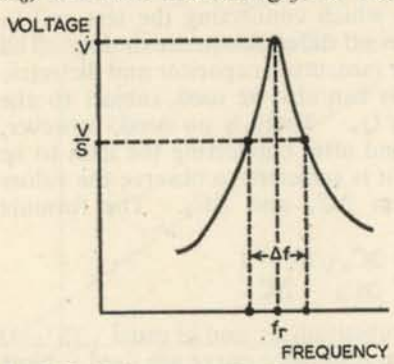


Fig. 9.28—Values measured in the frequency-variation method

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 reaction 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 two equations in that section become

$$Q_x = \frac{f_r C_x\sqrt{S^2-1}}{C_r(\Delta f_2 - \Delta f_1)} \quad \text{and } R_x = \frac{\sqrt{S^2-1}}{2\pi C_r(\Delta f_2 - \Delta f_1)} = \frac{2\pi f_r^2 L\sqrt{S^2-1}}{\Delta f_2 - \Delta f_1}$$

9.28. 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 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. 9.29 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

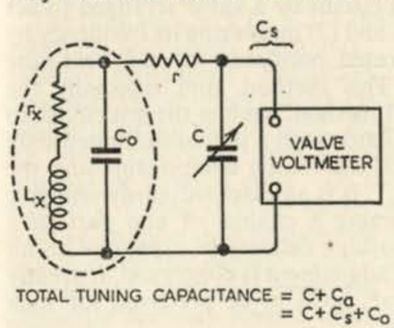


Fig. 9.29—Connections for the resistance-variation method of measuring r.f. resistance r_x; r is the resistance varied

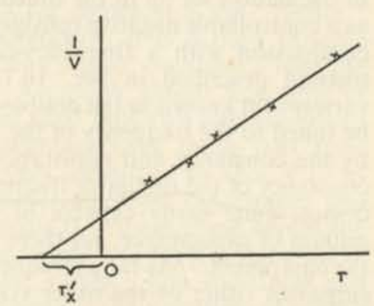


Fig. 9.30—Derivation of apparent inductor resistance (r_x') from readings of resonant voltage at a number of values of added resistance r in Fig. 9.29

* *J.I.E.E.*, November 1936; also *Proc. Wireless Section I.E.E.*, March 1937, pp. 6–18.

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. 9.30) 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. 9.27. 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-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. 9.26, 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.

9.29. Oscillator or Dynatron Measurements

In the third type of r.f. test equipment (Fig. 9.21*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.

* 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.

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. 9.21*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, $-10\,k\Omega$ or $-100\,k\Omega$, one might be inclined to say " $-100\,k\Omega$ ", and perhaps add " of course! " But bearing in mind that a parallel resistance of $10\,k\Omega$ represents much heavier losses than $100\,k\Omega$, one can see that $-10\,k\Omega$ 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 $+10\,k\Omega$ and $-100\,k\Omega$ is $+11{\cdot}1\,k\Omega$—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. 4.13, instead of resistance; but unfortunately many people are not used to thinking in terms of conductance.

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.

9.30. The Negative Resistor

The only item in Fig. 9.21*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. 4.13 to 4.15, 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 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 $-5\,k\Omega$, (conductance $-200\ \mu\mho$) others are limited to about -50 kΩ ($-20\ \mu\mho$). Two-valve circuits (Sec. 4.15) 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. The author has obtained oscillation from a Mazda AC/S2 at 60 Mc/s, but this seems to be quite exceptional.

Fig. 9.31 shows a practical dynatron circuit for measurement purposes. No particular valve is specified, as it depends on what is

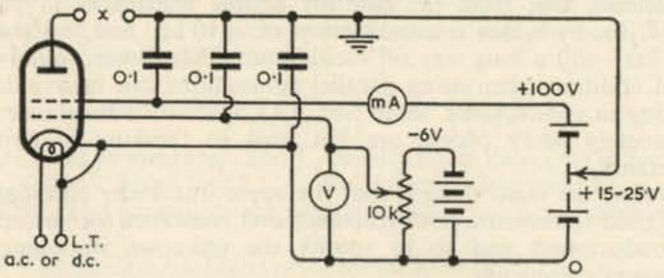


Fig. 9.31—Practical circuit of dynatron as used for measuring dynamic resistance, etc.

available (Sec. 4.13), 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. 4.5) 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, or even on, 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. Suitable circuit values for conductance control are shown in Fig. 4.13. In the following sections, " dynatron " includes transitron or any other negative resistance device.

9.31. Measurement of Dynamic Resistance

As with the other systems dealt with in Secs. 9.24 and 9.25, 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—the ordinary carbon type—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 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 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

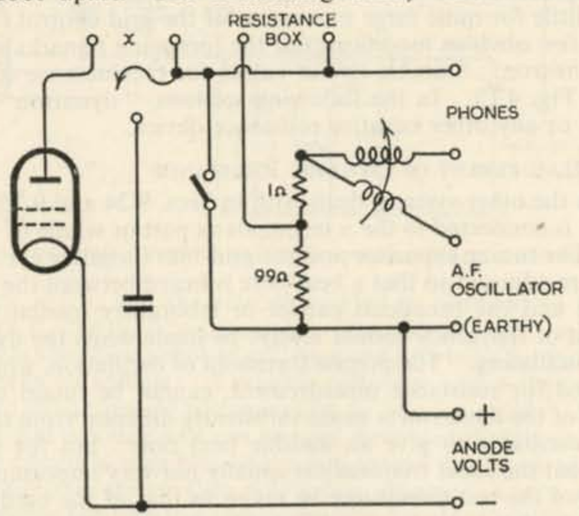


Fig. 9.32—Bridge circuit for measuring the negative resistance of a dynatron at any adjustment

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. 9.32 connected to a dynatron. 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 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. 9.33 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.

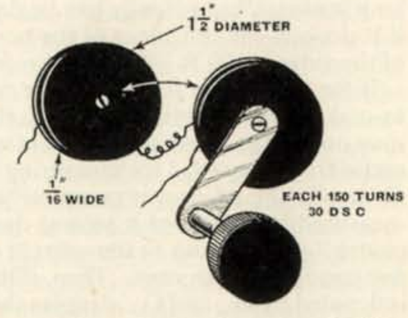


Fig. 9.33—Suitable dimensions for the phasing coils shown in Fig. 9.32

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 40 : 1 and 100 : 1, 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. 4.13).

Q, tan δ, 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. 9.26 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 $R_x=R_1R_2/(R_1-R_2)$.

9.32. 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. 9.26, 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, denoted by C_x, is connected to the x terminals in parallel with a 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 C_x set to its zero mark 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 per cent of f, and is therefore practically negligible compared with it, the formula can be simplified to

$$\Delta C_x \simeq \frac{101{\cdot}3\Delta f}{f^3 L} \qquad [\Delta f \text{ in kc/s; } f \text{ in Mc/s; L in } \mu\text{H; } \Delta C_x \text{ in pF}]$$

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. This is very 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.

9.33. 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 MEASUREMENT*

9.34. D.C. TESTS

As mentioned in the sections on valve-testing equipment (7.20 and 7.21) 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 much 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. 7.20.

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 μ_{g1g2}—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.

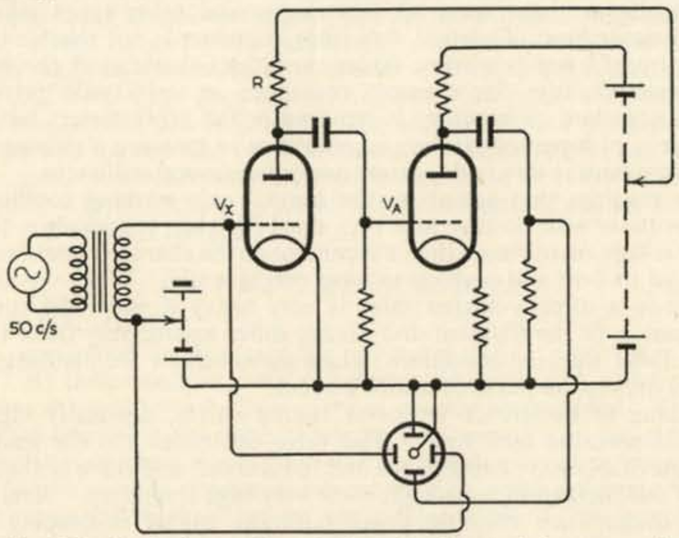


Fig. 9.34—Circuit diagram for projecting valve characteristic curves on the cathode-ray tube screen

9.35. 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. 9.34 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

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 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

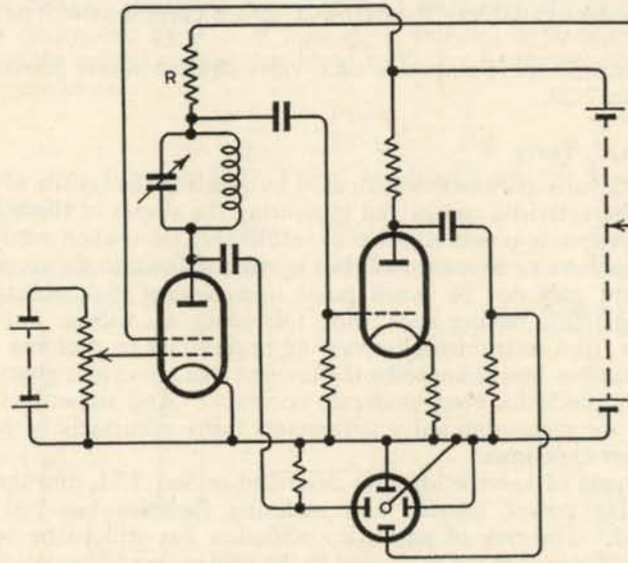


Fig. 9.35—Scheme of Fig. 9.34 modified for studying the oscillation characteristics of a dynatron

the cathode-ray apparatus is to show the behaviour of the valve under dynamic conditions. If R in Fig. 9.34 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

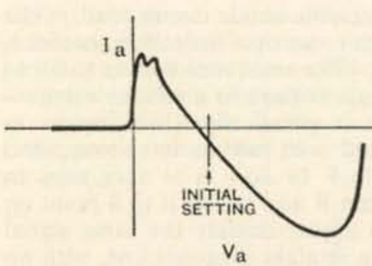


Fig. 9.36—Example of results obtained with the system of Fig. 9.35, showing over-oscillation due to insufficient negative grid bias

characteristic curve is swept over during oscillation; the anode current deflection is obtained as in Fig. 9.34, 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. 9.35 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. 9.36 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.

References to more elaborate c.r.t. valve equipment are given at the end of Sec. 7.20.

9.36. 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. 7.21, 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. 9.31, 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 in *Wireless World*, June 1945. It is analogous to the shunted-voltmeter method of measuring resistance (Sec. 9.1). Fig. 9.37*a* shows the equivalent circuit of a valve (Sec. 14.22) 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. 9.26 and 9.27, 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

$$\mu = \frac{V_1}{V_g}\left(\frac{r_a + R_L}{R_L}\right)$$

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

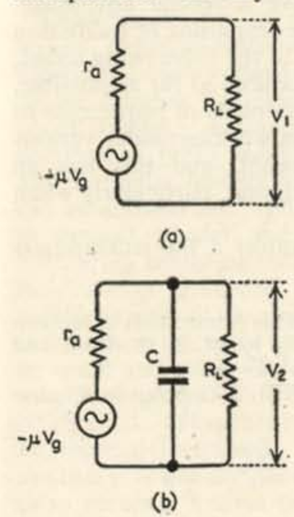


Fig. 9.37—Equivalent circuit of valve with input voltage V_g and load consisting of (a) resistance R_L, and (b) R_L shunted by capacitance C. From the output readings V_1 and V_2, r_a can be found

C should be large enough—say not less than 3 nF—to swamp any 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 $R_L = 10\ k\Omega$ and $C = 0{\cdot}1\ \mu F$. It should be noted that if negative feedback is in use the r_a indicated is the value as modified thereby.

9.37. 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. 14.20) 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. 9.24 to 9.32,† but as the capacitances are usually small the oscillator methods (Sec. 9.29) 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.

* 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.*, August 1938, pp. 1011–32.

† For details of measurement at 1·5–300 Mc/s, see M. J. O. Strutt in *Wireless Engineer*, September 1937, pp. 478–488.

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. 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 measure 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. 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 impedance by valve voltmeter, in which

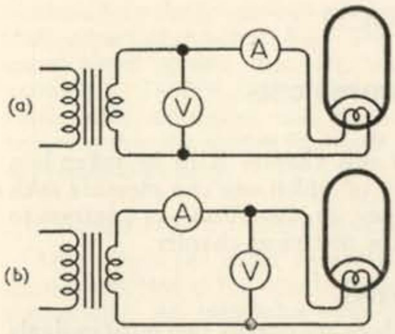


Fig. 10.1—Which method of connection is preferred depends on whether it is easier to allow for (a) the voltage drop in the ammeter, or (b) the current taken by the voltmeter. In low voltage a.c. circuits like this, b is usually the better

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 must only be used with caution when measuring voltages in valve circuits. For the same reason it is unsuitable for use with an a.c. valve or a metal-rectifier voltmeter

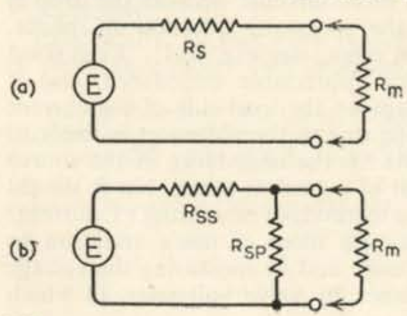


Fig. 10.2—The reduction in voltage between two points caused by connecting a voltmeter (resistance R_m) can easily be calculated if the source of voltage is equivalent (a) to an e.m.f. in series with a known resistance R_s. In the slightly more complicated case b, Thévenin's theorem shows the source 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

anywhere near the non-linear part of its characteristic, where its impedance varies rapidly with voltage.

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. 5.12).

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 instrument, 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. 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;

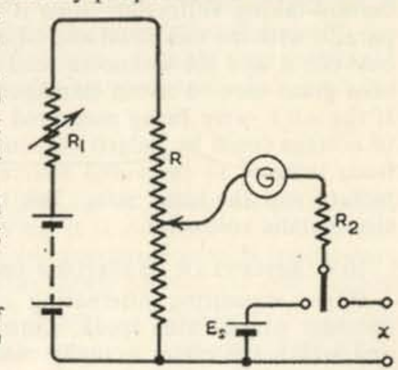


Fig. 10.3—Diagram showing principle of potentiometer, used to compare an unknown voltage at x with the standard E_s

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 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.

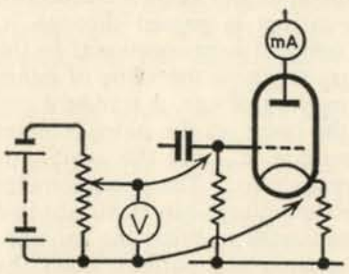


Fig. 10.4—Example of how to use a current-operated voltmeter V to measure a voltage without drawing any current

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. 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 5.11). 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 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.

10.5. 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

* "Total Power." *Wireless World*, March 1952, pp. 117–120.

† "Noise." *Wireless World*, May and June 1952, pp. 199–202 and 222–6.

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.

10.6. Phase Difference

The cathode-ray method of measuring phase difference, described in Sec. 9.9 in connection with capacitors, is of general application. 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. 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 in *Electronic Engineering*, June 1950, pp. 238–242.

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.

A more precise and flexible method is by a.c. potentiometer (Sec. 10.2). Special instruments have been produced for accurate measurement of phase, particularly in connection with negative feedback and servo mechanisms, but these are elaborate and costly and outside the scope of this book.

10.7. 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, with the respective Y and X plates, to the c.r.t. anode, and preferably also to earth. These connections are shown diagrammatically in

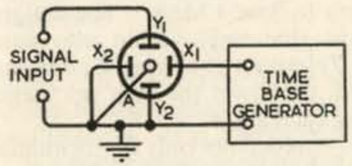


Fig. 10.5—Normal method of connecting a cathode-ray tube for examining signal waveform from a source having one terminal earthy

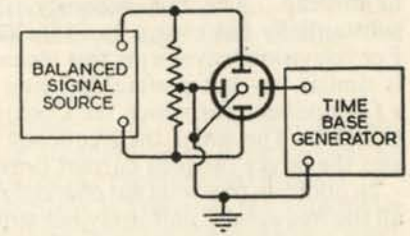


Fig. 10.6—Modification of Fig. 10.5 to suit a balanced source which has no centre or earth terminal

Fig. 10.5. Usually shift voltages are introduced between anode and the plates X_2 and Y_2. With the Cossor double-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. 10.6.

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, 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 connect a cathode follower as close up to the actual source as possible. The output impedance of a cathode follower is only a few hundred ohms, so it allows for a yard or two of low-capacitance connecting cable as well as the capacitance of the Y plates, etc. If worked under suitable conditions a cathode follower reproduces the input (grid) waveform at the output (cathode) with negligible distortion, thanks to its 100-per-cent negative feedback; but if it is to handle waveforms like television pulses without either serious distortion or limitation of amplitude the load resistance must be quite low—of the order of two or three thousand ohms.*

The voltage output of a cathode follower is no greater than its input—in fact slightly less—so when the signal voltage is inadequate, or when a current waveform is to be studied (in terms of the voltage drop across a series resistance), an amplifier is needed, and obviously if it introduces appreciable distortion of any kind the results are liable to mislead. The first necessity is a frequency characteristic that is substantially flat over a band including all the frequencies concerned. For television waveforms this means zero to 3 or 4 Mc/s. The design is similar to—but preferably better than—that employed in television v.f. amplifiers, for which see Cocking's *Television Receiving Equipment* (Iliffe).† The wider the frequency band, the lower the gain per stage and the larger the feed current drawn by each stage.

In addition to a wide flat characteristic—needed not only for including all the frequencies uniformly but still more for avoiding phase distortion—there are other requirements such as linearity and absence of transient distortion.

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 in *Wireless Engineer*, September 1951, pp. 261–5, and a more elaborate one by R. Anderson and J. R. Smith in *Proc. I.E.E.*, Pt. IIIA, No. 19 (1952), pp. 591–6.

10.8. Waveform Analysis

An oscilloscope gives a general picture of waveform, but for

* "Cathode-follower Dangers", by W. T. Cocking. *Wireless World*, March 1946, pp. 79–82.

† A very suitable amplifier covering 8 c/s to 8 Mc/s, linear up to 50 V, is described by V. H. Attree in *Electronic Engineering*, November 1952, pp. 504–6.

quantitative results the most useful instrument is a wave analyser. It enables the amplitudes and frequencies of all harmonics and any other constituents (such as intermodulation products) to be measured. It can be used not only on signals of a definite frequency but for noise.* Commercial wave analysers, described in Sec. 5.20, are essentially highly selective valve voltmeters, and are rather costly. Though less convenient in use, a method of wave analysis described by D. Martineau Tombs in *Wireless Engineer*, July 1950, pp. 197–200, is much less expensive, 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. 9.26). 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.

If a really accurate picture of a waveform is available, the amplitudes of a limited number of its harmonics can be calculated by a procedure based on its instantaneous values at selected phase angles. These methods are explained in some of the larger books on a.c., and by R. C. de Holzer in *Electronic Engineering*, June and July 1945, and D. R. Turner in the same journal, January 1953. See also Sec. 11.14 for a related method based on the non-linearity of the cause instead of the waveform itself.

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. Now we come to consider the methods of comparison.

There are two main classes of method, according to whether the indicator is visual or aural. 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. In general one can say that visual (chiefly cathode-ray) methods of frequency comparison are best for low frequencies and aural methods for high frequencies.

The very highly accurate standards, with errors of one or less in a

* For example, "Gramophone Record Scratch", by M. G. Scroggie. *Wireless World*, November 1939.

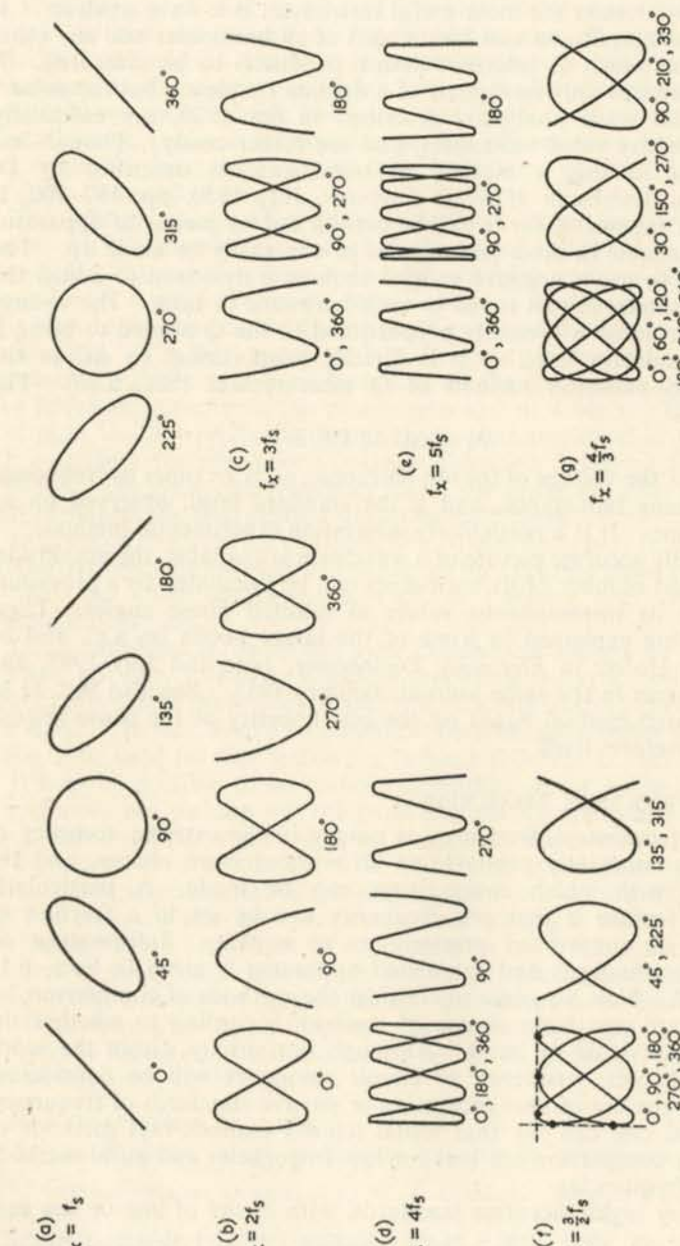


Fig. 10.7—Examples of Lissajous figures obtained on a c.r. tube when sinusoidal voltages having the frequency ratios and phase differences specified are applied to the two pairs of plates

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 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 (see Sec. 6.14). 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. 10.7*a*; see also Sec. 9.9. Throughout Fig. 10.7 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. 10.7*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. 10.7*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. 10.8. 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 toothedwheel pattern, in which the number of teeth indicates the frequency ratio, and a slight departure from exact ratio makes the wheel rotate. 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. 10.9*a*) would indicate exactly 50 c/s.

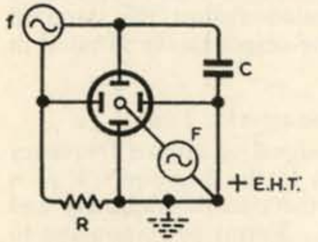


Fig. 10.8—Method of connecting two signal sources f and F so as to obtain traces such as Fig. 10.9, in which there are F/f teeth to the wheel. The reactance of C at frequency f should be approximately equal to R

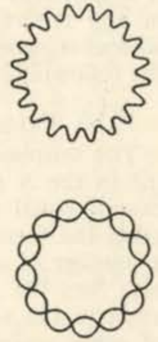


Fig. 10.9—Two traces obtained by the method shown in Fig. 10.8. Example a shows F = 20f; b, F = 13f/2

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. 10.9*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, $16\frac{2}{3}$, $12\frac{1}{2}$, 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. However, when the system is not overloaded it is usually well within 1 per cent 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

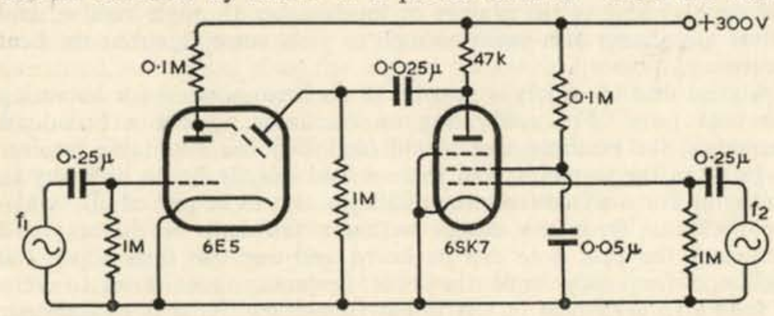


Fig. 10.10—Circuit of indicator for showing when f_2 is equal to f_1 or a multiple thereof up to about 10

possible arrangements; the one shown in Fig. 10.10 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

* *Electronic Engineering*, May 1951, p. 197. Another circuit, by H. V. Beck, appears on p. 405 of the October 1951 issue of that journal.

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. 9.27), it is desired to establish 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 per cent, that would represent an uncertainty of only a little over 0·01 per cent 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. *Electronic Engineering*, April 1947, pp. 134–5.

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 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. 10.11 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

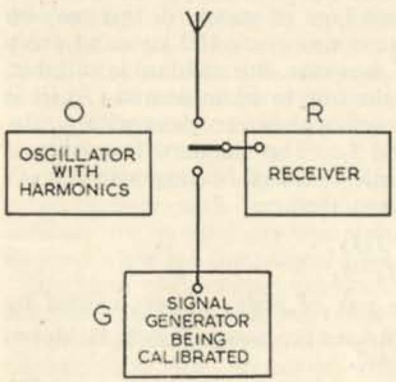


Fig. 10.11—Method of calibrating the frequency scale of a signal generator from a single known radio frequency

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. 12.7.

10.15. Syntonizing Passive 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.

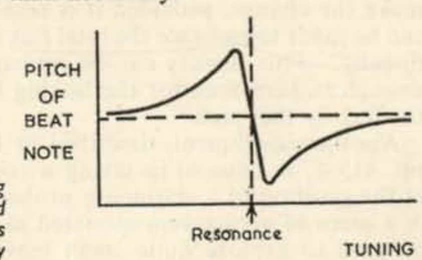


Fig. 10.12—" N curve " of the disturbing effect of a loosely-coupled circuit tuned through resonance—a phenomenon that is helpful in syntonizing circuits very precisely

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. 10.12. Exact 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 in *Wireless Engineer*, December 1931, pp. 639–646.

10.16. Measurement of 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 no longer be omitted altogether from even a book like this. A good account of the established methods and equipment is given in F. G. Spreadbury's *Permanent Magnets*, Ch. VII (Pitman). There is room here only for two recent developments.

The usual 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 in *Wireless World* (April 1951, pp. 161–2), which is no more than special type of valve voltmeter, readily made from parts likely to be found in the radio 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 0–1 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.

Another instrument, described in *Wireless World*, November 1950, pp. 415–6, is unusual in giving a continuous reading of flux density at the position of a stationary probe. It makes use of the Hall effect in a piece of germanium mounted at the end of the probe, which can be used to explore quite small magnetic gaps. Without going into

Fig. 10.13—B.T.-H. direct-reading fluxmeter, employing the Hall effect in germanium. The point of the probe is held in the magnetic field to be measured

detail, one can say that the Hall effect is a distorting of the lines of current flow in a conductor placed in a magnetic field, whereby a measurable p.d. is set up between points at each side of the conductor. An instrument of this type made by the B.T.-H. Co., and shown in Fig. 10.13, has three ranges and a calibration-checking facility. The full-scale readings are 5, 10 and 25 kilogauss: in m.k.s. units, 0·5, 1 and 2·5 Wb/m^2.

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. 9.23 to 9.31, 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, a frame aerial can be regarded as a large coil.

11.2. TRANSMISSION LINES OR FEEDERS

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, 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,

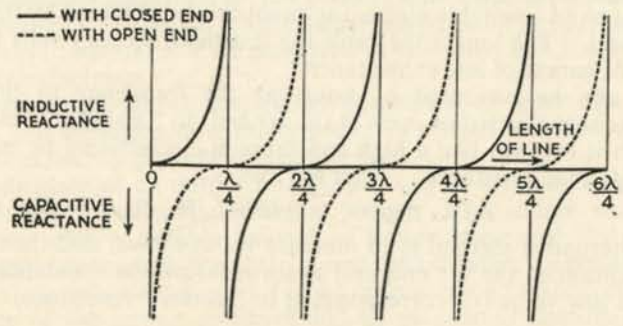


Fig. 11.1—Showing how the reactance of short-circuited and open-circuited lines varies with length

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. 12.9), 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, and the loss per 100 ft is usually 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

$$Z_0 = 1/2\pi f\sqrt{C_s C_p}$$

As a guide, at frequencies in the British television band 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

$$Z_0/R_x \text{ nepers}, = 8{\cdot}686\ Z_0/R_x \text{ db} \qquad \text{(Sec. 14.34)}$$

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:

$$Z_0 = \sqrt{r_x R_x}$$

and the loss is $\sqrt{(r_x/R_x)}$ nepers or $8{\cdot}686\sqrt{(r_x/R_x)}$ 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

$$v = 2l\,(f_2 - f_1)$$

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 (Sec. 14.16 and Fig. 14.23). 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. 9.16 to 9.20.

Leakage inductance is relatively small, and is a measure of the extent to which the primary is not completely coupled to the secondary. It 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.

Self-capacitance is an important parameter of intervalve transformers, but these are now rarely used; in output transformers self-capacitance usually plays only a minor part.

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. The method is not recommended unless the efficiency of the output-meter transformer can be relied upon to be reasonably equal at all the frequencies concerned.

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. *Wireless Engineer*, September to November 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. 9.16 to 9.20 for reactance and resistance of coils, particularly those with iron cores. D. E. L. Shorter has shown (*Wireless World*, November 1950, p. 383) 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 loudspeaker 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 the most important—notably the frequency characteristic—call for not only a very high-quality microphone which must be calibrated, but also acoustically dead surroundings. This technique is outside the scope of this book, but here are a few references to sources of information:

Acoustic Measurements, by L. L. Beranek (Chapman & Hall, 1949).
Loud Speakers, by N. W. McLachlan (O.U.P., 1934).
" Loudspeaker Response Curves ", *Wireless World*, February 1952, pp. 58–9, 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

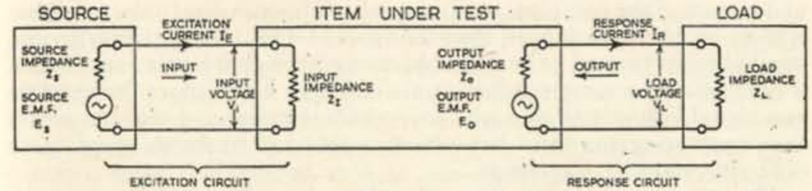


Fig. 11.2—Diagram showing standard terms and symbols used in connection with tests on amplifiers, receivers, etc.

have been standardized (B.S.2065 : 1954) and these are shown in Fig. 11.2, 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. 14.22).

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 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 invariably 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 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. 11.3 shows the latter arrangement. The terminations

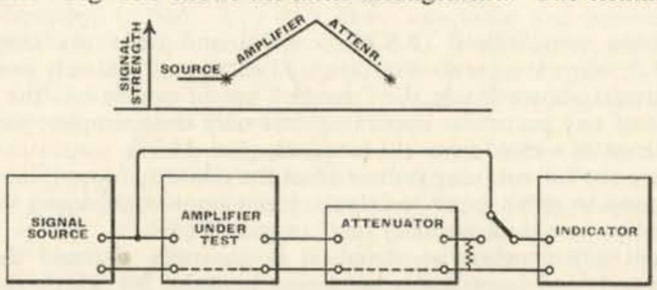


Fig. 11.3—Block diagram of the arrangement of apparatus when measuring the gain of any amplifier by comparing it with the loss introduced by a calibrated attenuator. If the attenuator is continuously variable over the range concerned, the only purpose of the meter is to adjust the signal level to equality 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

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. 11.3 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 against one another 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 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

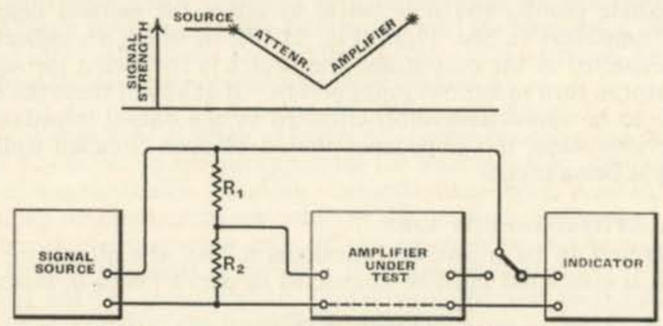


Fig. 11.4—It may be better in some cases for the amplifier and attenuator to be at low signal levels, source and meter being high. This is done by putting the attenuator ahead. This diagram also shows the use of a simple fixed or semi-variable attenuator to bring the meter points to approximate equality. A meter calibration is then needed as well as the attenuator calibration, but the need for wide range is avoided

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. 11.4) 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. 10.7*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, with ink or otherwise.

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. 11.17), 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

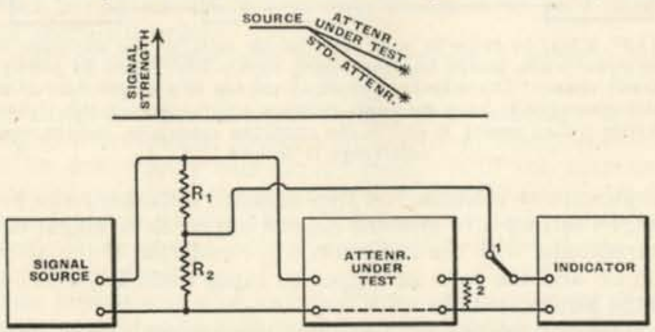


Fig. 11.5—Here two attenuators are being compared, and the only difference in method is that they are connected in parallel instead of series. If possible the standard is adjusted to make the two stars coincide

in cascade; otherwise the procedure is the same. Fig. 11.5 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, obtainable from science stationers, having a logarithmic horizontal scale for frequency and a linear vertical scale for db. 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 set up 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. An example of a rapid curve-tracing system is described by W. N. Weedon, in *Wireless World*, 13 August 1937, p. 137, and by the present author in the same journal, 11 May 1939, p. 435. The principles of the subject are discussed by D. W. Thomasson in *J. Brit. I.R.E.*, 1948, pp. 171–188.

A portable a.f. curve tracer by Industrial Electronics consists of an oscilloscope in which the Y deflection is obtained from the output of the unit under test, and the X deflection from a RC discriminating circuit giving a voltage proportional to the logarithm of the signal-source frequency, which is swept continuously over the a.f. band.

11.9. Avoidance of Overloading

In the foregoing descriptions of gain and frequency characteristic measurements, the only references to signal amplitude have been

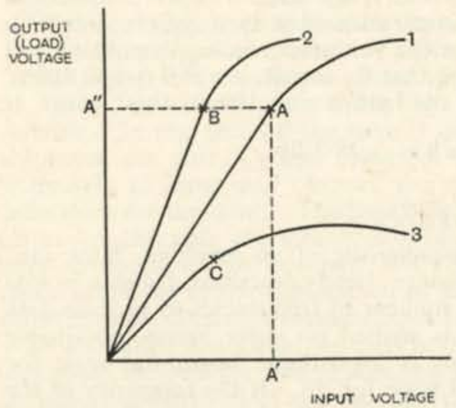


Fig. 11.6—Showing how signal level that is within the capabilities of an amplifier in one set of conditions may overload it in another

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. 11.6. 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 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. 11.4 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. 11.7. 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.

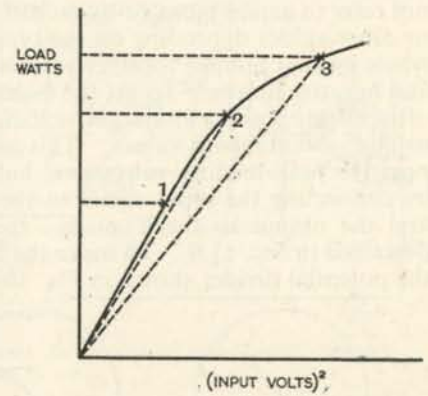


Fig. 11.7—Typical input/output characteristic of an amplifier or other unit, showing how estimation of the overload point is a matter for judgment

Instruments for measuring output power are discussed in Sec. 5.7. 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 be considerably different 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. 11.7 rather indefinite; the curve is unsuitable as a basis for more precise analysis of non-linearity. The points on it do 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. 11.4 will probably have to be used.

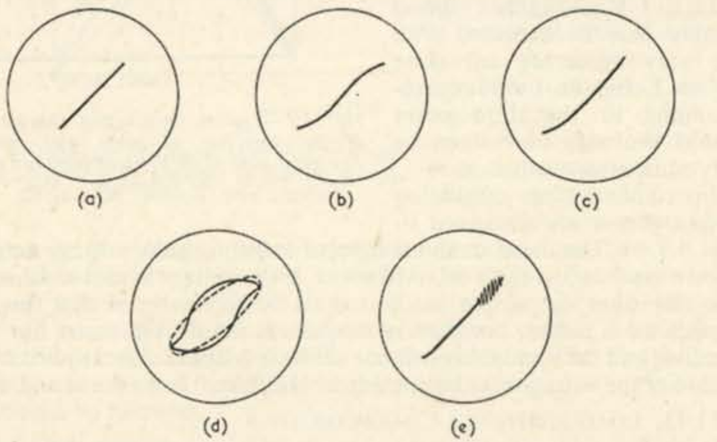


Fig. 11.8—Representative c.r. tube traces when input signal is applied to X plates and output to Y, showing the appearance of (b and c) non-linearity, (d) phase shift and (e) parasitic oscillation

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. 11.8*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. 9.9.

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

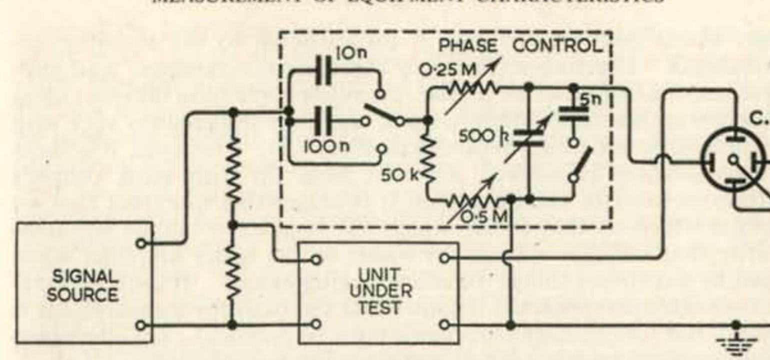


Fig. 11.9—Suggested phase-shift network for facilitating non-linearity measurements by c.r.t.

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. 11.8*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. 11.8*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. 11.9 are likely to cover most requirements.

Fig. 11.8*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 half an hour than days of testing with meters.

11.13. Basis of Non-linearity Measurements

Although the test just described can give so clear a picture of the existence of non-linearity, and a good idea of its magnitude and character, one cannot be satisfied without an actual measurement. 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.

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 per cent ninth harmonic sounds worse than 10 per cent 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

* *Proc. Wireless Section I.E.E.*, September 1937, p. 179.

† " The Influence of High-order Products in Non-linear Distortion." *Electronic Engineering*, April 1950, pp. 152–3.

‡ *Wireless World*, 19 May 1938, pp. 440–2.

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. 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. 11.8*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 following are some references to add to those already given:

" Intermodulation in A.F. Amplifiers ", by A. C. Bartlett. *Wireless Engineer*, February 1935, pp. 70–74.

" Amplitude Distortion ", by J. H. O. Harries. *Wireless Engineer*, February 1937, pp. 63–72.

" An Analysis of the Intermodulation Method of Distortion Measurement ", by W. J. Warren and W. R. Hewlett. *Proc. I.R.E.*, April 1948, pp. 457–466.

" Intermodulation Distortion ", by Thomas Roddam. *Wireless World*, April 1950, pp. 122–5.

" Intermodulation Distortion ", by A. P. G. Peterson. *General Radio Experimenter*, March 1951, pp. 1–8.

" Intermodulation Testing ", by E. W. Berth-Jones. *Wireless World*, June 1951, pp. 233–6.

" Relations Between Amplitudes of Harmonics and Intermodulation Frequencies ", by M. V. Callendar. *Electronic Engineering*, June 1951, pp. 230–2, corrected in the issue of July 1951, p. 277.

" Simple Distortion Meter ", by V. J. Tyler. *Wireless World*, September 1953, pp. 431–3.

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

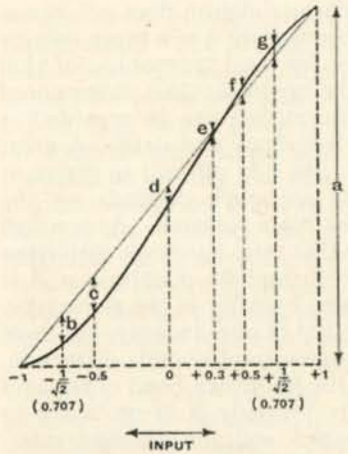


Fig. 11.10—Typical transfer characteristic, showing dimensions needed for harmonic analysis as described

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. 11.9.

The following method is adapted from that described by J. A. Hutcheson (*Electronics*, January 1936) who demonstrates it for all harmonics up to and including the seventh; and it involves only simple arithmetic. Fig. 11.10 shows how the requisite data are derived from the cathode-ray trace. The total output swing, a, is measured, which can more 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, 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.)

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.

As an example, suppose that in Fig. 11.10

$$a = 120$$

$$b = -10 \qquad e = -2$$
$$c = -12 \qquad f = 2$$
$$d = -7 \qquad g = 3$$

Then $$V_2 = \frac{2-12}{3} + \frac{-7+10-3}{4} = -3{\cdot}33$$

$$V_3 = \frac{14}{3} = 4{\cdot}67$$

$$V_4 = 0$$

$$V_5 = 4{\cdot}67 + \frac{-10-3}{2{\cdot}828} = 4{\cdot}67 - 4{\cdot}60 = 0{\cdot}07$$

$$V_6 = -\frac{7}{2} + 3{\cdot}33 = -0{\cdot}17$$

$$V_7 = \frac{-2+(1{\cdot}82\times 3{\cdot}33)-(1{\cdot}092\times 4{\cdot}67)-(0{\cdot}699\times 0{\cdot}07)+(0{\cdot}751\times 0{\cdot}17)}{1{\cdot}146}$$

$$= \frac{-2+6{\cdot}07-5{\cdot}10-0{\cdot}05+0{\cdot}13}{1{\cdot}146} = -0{\cdot}83$$

$$V_1 = 60+4{\cdot}67-0{\cdot}07-0{\cdot}83 = 63{\cdot}77$$

	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

For calculation of harmonics from waveforms, see references in Sec. 10.8.

11.15. Measurement of Harmonics

There are instruments and methods for measuring total harmonic distortion by using a filter or bridge to remove the fundamental component from the output, but as the result does not give even an approximate measure of subjective distortion no further attention will be given to these here. To obtain material for a fair assessment it is necessary to find the amplitudes of all significant harmonics—bearing in mind that small fractions of 1 per cent 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. 5.20). 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 high-selective valve-voltmeter type is so expensive that various alternatives have been proposed. One of these is described in Sec. 10.8. In another, beats slow enough to observe on a d.c. meter are set up between an exploring oscillator and each harmonic in turn; but extreme frequency stability is essential for any hope of success.

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 (see reference to J. H. O. Harries on p. 301). 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

* This is sometimes called first-order intermodulation.

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 (see reference to A. P. G. Peterson on p. 301), 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 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

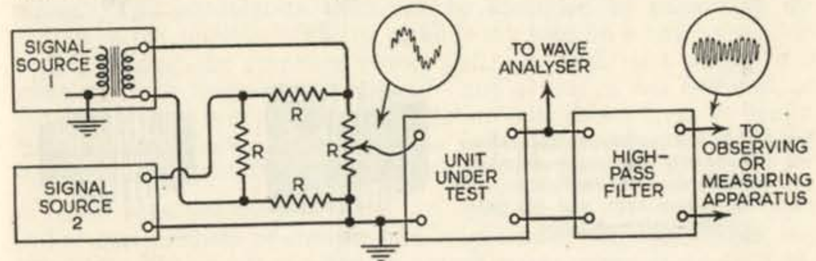


Fig. 11.11—Method of connecting two a.f. signal generators for intermodulation tests. The R bridge is used to prevent the generators from affecting one another

Fig. 11.11 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.

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, as it were, through a microscope. That is to say, if it is small it can be amplified. Fig. 11.12 shows the simple requirements

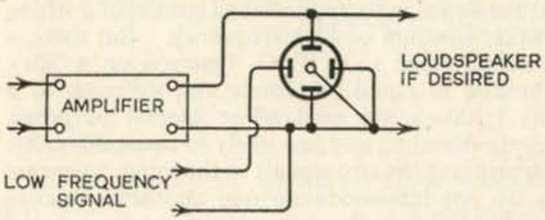


Fig. 11.12—Connection of c.r.t. for observing intermodulation. Typical traces are shown in Fig. 11.13

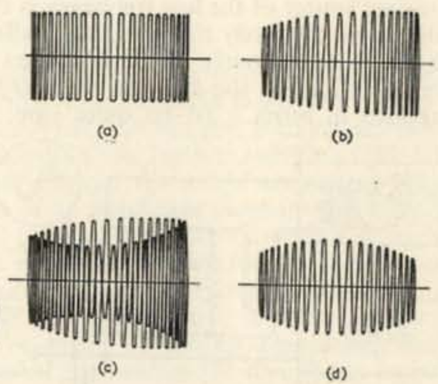


Fig. 11.13—Intermodulation traces showing (a) linearity (no intermodulation), (b) second-order intermodulation, (c) same with phase shift, and (d) third-order intermodulation

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. 11.13*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 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 per cent. Fuller information is given in the papers referred to on p. 301, and design data for a simple intermodulation meter is given by J. M. van Beuren in *Audio Engineering*, November 1950, pp. 24–25 and 56–58.

Intermodulation distortion is usually several times as great as the measured harmonic distortion, as one would expect when it is a percentage of the high-frequency signal whereas harmonics are a percentage of the stronger low-frequency signal.

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 in reducing 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 relatively 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. 5.20 (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.

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 an almost infinite range of frequency 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. 11.14

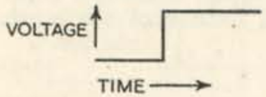


Fig. 11.14—Step voltage, which is the basis of transient tests

(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. 4.31.

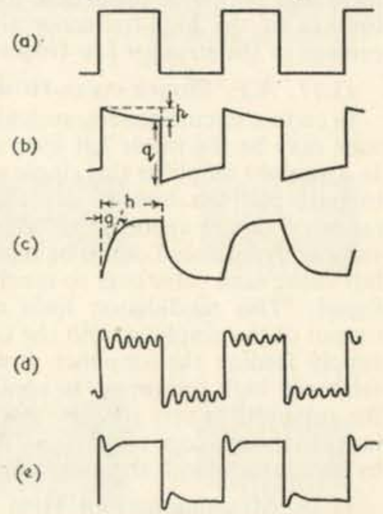


Fig. 11.15—Characteristic traces obtained in square-wave tests, from which the performance of the amplifier under test can be inferred

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 a perfect square wave (Fig. 11.15*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 the usual type of capacitance coupling. Assuming that the droop, measured by p/q, is not much more than 0·3, and denoting the signal frequency by f_s, and the frequency at which the coupling attenuates to the extent of 3 db by f_1, then

$$f_1 \simeq 2\,p\,f_s/\pi\,q$$

A suitable f_s for testing typical a.f. amplifiers is therefore about 150–300 c/s.

The phase lag due to falling-off at the high frequencies shows up as at *c*. Assuming the cause is 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_2, can be found by measuring the *initial* rate of rise, as shown; since

$$f_2 \simeq h\,f_s/\pi\,g$$

For g to be measurable, without conflicting with the assumption, f_s for this test ought to be about 0·3 to 1 times f_2—say 5,000 c/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.

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 4.1. 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 in *Wireless World*, July 1951, pp. 256–9, and discussed by L. J. Elliott in the September 1951 issue of that journal, p. 370.

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. 4.20). 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 shown in Fig. 11.16 and described by V. H. Attree in *Electronic Engineering*, March 1949, p. 100.

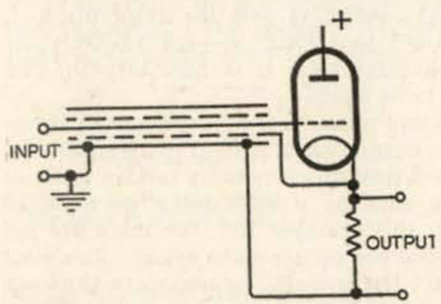


Fig. 11.16—Cathode-follower device for reducing the input capacitance of a screened lead

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 in *Electronic Engineering*, June 1949, pp. 204–8; for more detailed analysis of the v.f. amplifier by this method, see P. M. Seal in *Proc. I.R.E.*, January 1949, pp. 48–58; and for general television receiver testing, see M. V. Callendar in *Wireless World*, February 1952, pp. 42–5.

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 is seldom, however, 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

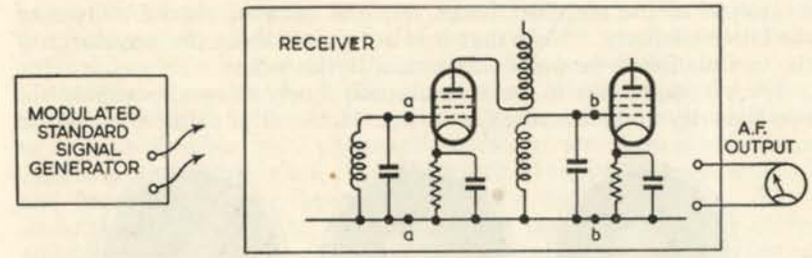


Fig. 11.17—Method of measuring the gain of an r.f. stage, by applying a signal generator in turn to aa and bb

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 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. 11.17 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. 11.18*a*. The depth of modulation is 100 (A − B)/(A + B) per cent, 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

Fig. 11.18—C.r.t. traces from which the purity and depth of modulation can be determined

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. 9.24 to 9.31; 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 popular 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 per cent 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,

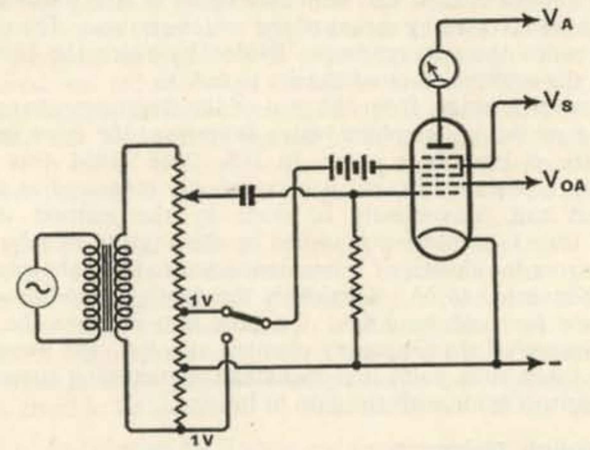


Fig. 11.19—Circuit used for measuring the conversion conductance of a frequency-changer valve, employing one low-frequency source only

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

* " Diode-detector Distortion ", by W. T. Cocking. *Wireless World*, May 1951, pp. 171–2.

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, in *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. 11.19 shows the arrangement applied to a heptode valve. The electrodes are fed with their appropriate voltages, and the oscillator 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 1 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 a complete stage, from the grid of the frequency-changer valve to the grid of the i.f. amplifier valve following, the most important 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. 11.17 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, need not extend to great length in spite of its importance, 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.

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

* "R.M.A. Specification for Testing and Expressing Overall Performance of Radio Receivers". *Proc. Wireless Section I.E.E.*, September 1937, pp. 179–197.

Standards on Radio Receivers, 1938. (The Institute of Radio Engineers, U.S.A.).

"Proposed Test Procedure for F.M. Broadcast Receivers", by D. Maurice, G. F. Newell, and J. G. Spencer. *Electronic Engineering*, March 1952, pp. 106–111.

Recommended Methods of Measurement on Receivers for Amplitude-Modulation Broadcast Transmissions (Publication 69), Central Office of the International Electrotechnical Commission, Geneva, Switzerland (1954). Obtainable through B.S.I. (see p. 11).

See also "The Testing of Communication-type Radio Receivers", by W. J. Bray and W. R. H. Lowry. *J. I.E.E.*, Pt. IIIA, 1947, pp. 313–327. Discussion: *J. I.E.E.*, Pt. III, 1948, pp. 271–6.

the measured characteristics, therefore, 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 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. 4.25 shows the specification of the standard dummy aerial connected

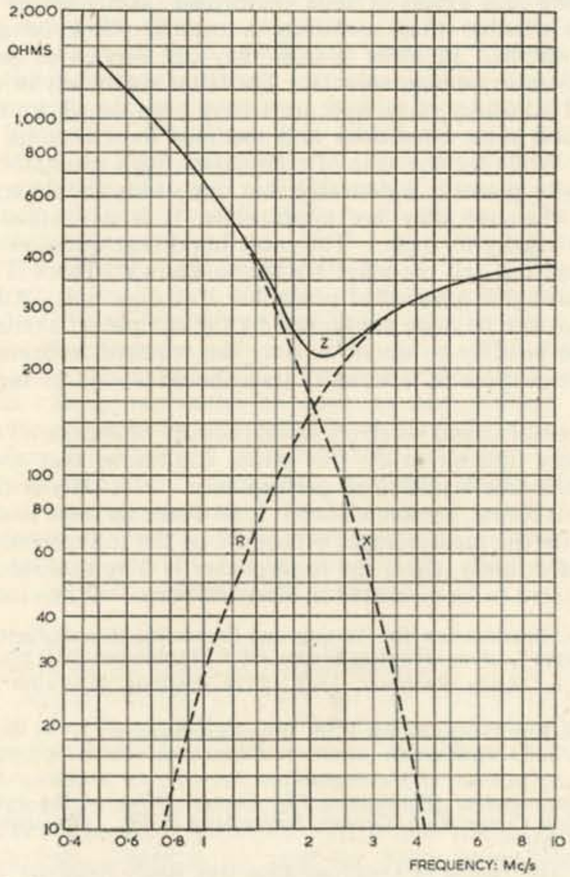


Fig. 11.20—Series resistance, reactance, and magnitude of impedance of standard dummy aerial (Fig. 4.25) as a function of frequency

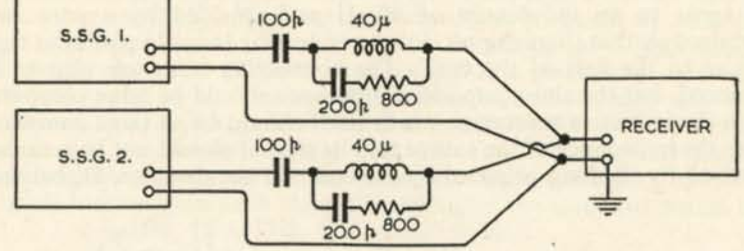


Fig. 11.21—Method of connecting two r.f. signal generators for selectivity measurements. Note that the dummy aerials have twice the standard impedance, and the generators have to be set to give twice the desired microvoltage

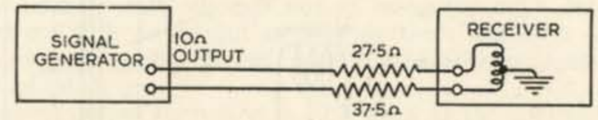


Fig. 11.22—Method of connecting unbalanced signal-generator output to balanced input

between signal generator and aerial terminal when testing broadcast sound receivers, and Fig. 11.20 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. 11.21, 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 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. 11.22.

Injection of the signal into a frame 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. 11.23). 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 frame aerial and D cm away from it along that axis, then the field strength in microvolts per metre at the frame aerial when the signal generator output is V microvolts is

$$\frac{18{,}850\ N\ R^2\ V}{[D^2 + (H/2)^2]^{3/2}\ X} \qquad [D \nless \text{double } H, W \text{ or } 2R]$$

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 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

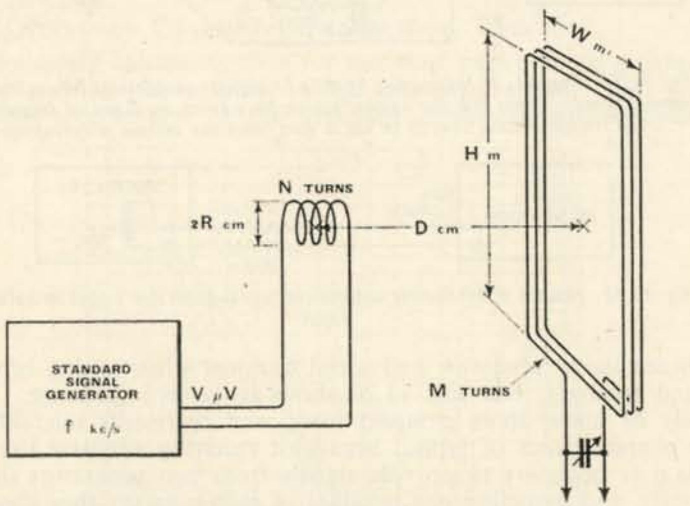


Fig. 11.23—Preferred method of introducing a known signal into a frame aerial. The signal strength is calculated from the quantities shown

centres of this coil and the frame 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 frame aerial, the field strength obtained as above must be multiplied by the effective height of the aerial, which for a rectangular frame is approximately

$$2\ M\ H \sin \frac{\pi f W}{300{,}000}\ \text{metres}$$

where M is the number of turns
H the height in metres
W the width in metres.

The voltage developed across the aerial at resonance is, of course, Q times the voltage induced as described.

The standard modulation frequency and depth are 400 c/s and 30 per cent 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, so it is advisable 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 per cent 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-impendance coupling coil were used. Of course, 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 only suffices 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.

Suppose that for the purpose in view sensitivity is defined as the minimum r.f. input voltage modulated 30 per cent 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 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

* J. M. Pettit in *Proc. I.R.E.*, March 1947, pp. 302–6.

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.

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; then, 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 further reduction is expressed as its 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. 4.32), the maker's instructions for which should be followed. Information can also be found in the paper by Bray and Lowry (footnote on p. 315) and in " The Theory and Measurement of Noise Factor ", by R. J. Yates in *Marconi Instrumentation*, July to November 1950.

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 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. 11.24).

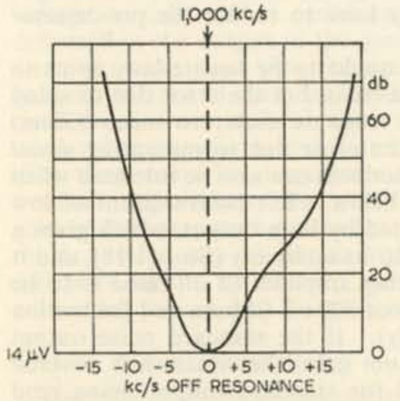


Fig. 11.24—One method of presenting the results of a single-generator selectivity test. Reckoning the input at resonance as zero db facilitates comparison with other results. The sensitivity is also shown, on the left

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 ± 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.

11.33. Two-signal Tests

The two signal generators are connected as in Fig. 11.21 (remember to use double-impedance dummy aerials and double the desired signal voltages!). 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 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

* "The Mutual Interference of Wireless Signals in Simultaneous Detection", by E. V. Appleton and D. Boohariwalla. *Wireless Engineer*, March 1932, pp. 136–9.

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. 11.24), 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.

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.

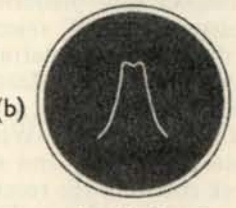
The measurement of hum itself is described in Sec. 11.18, but faulty design may give rise to an increase when a carrier wave is present, due to modulation of the carrier at hum frequency. So when testing a whole receiver the hum output should be measured with an unmodulated carrier wave tuned in, at 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. Instantaneous Selectivity-curve Tracing

When investigating or aligning selective circuits, especially those intended to give a broad flat top and steep sides, the plotting of selectivity curves such as Fig. 11.24 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. 4.29 and 5.29. It consists of a signal generator continuously swept over the required band of frequency, synchronously with the time base of an oscilloscope. The set-up shown in Fig. 5.36 is basically the same whether the generator is wobbulated mechanically (as shown) or electronically (as is more usual). The input connection

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. 11.25*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

(a)



(b)

Fig. 11.25—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

amplification is needed; at 465-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. 11.25*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.

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 465 kc/s. The oscillator will then be working at 1,465 kc/s, the image frequency being 1,465 + 465 = 1,930 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., 930 and 1,395 kc/s if the i.f. is 465 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. 315).

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 is an example (Fig. 11.26*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

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

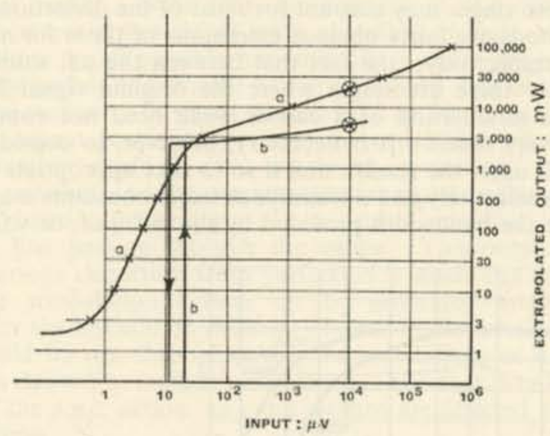


Fig. 11.26—Example of a.g.c. characteristics. Curve a is taken with no muting system in operation. At the extreme left a noise output of 2 mW with no signal is shown; then there is a steep rise during which a.g.c. is inoperative because the bias voltage has not been overcome. When it is, a.g.c. restricts further rise. Owing to excessive bias, the flat top is too high. With muting in use, curve b results

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 per cent modulation this should normally be slightly greater than the rated maximum output. In Fig. 11.26*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 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. 11.26*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

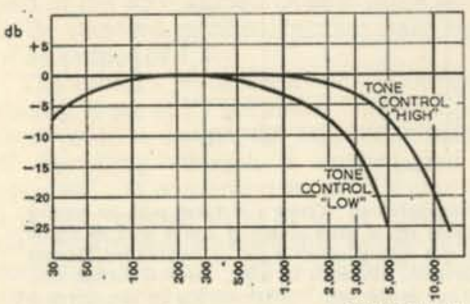


Fig. 11.27—Typical overall modulation-frequency response curve of a receiver

alone. Fig. 11.27 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 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 per cent 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 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, mains 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

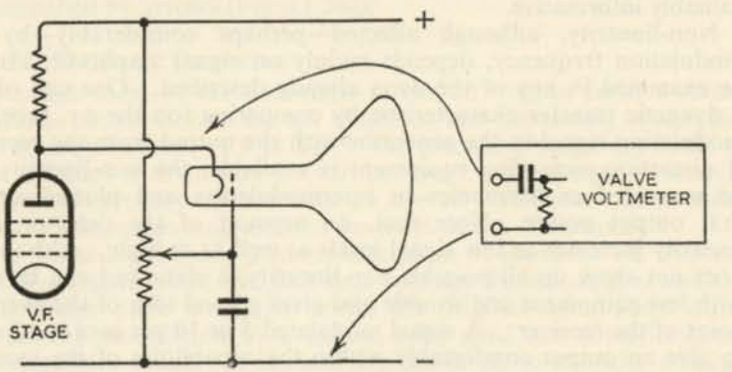


Fig. 11.28—Method of connecting a valve voltmeter to read the output of a vision channel. The voltmeter must be provided with a low-capacitance probe

range of voltages at which the set is rated to work on any one mains-voltage tapping.

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

the output is read as voltage at the control electrode of the c.r.t. Fig. 11.28 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. 5.12).

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

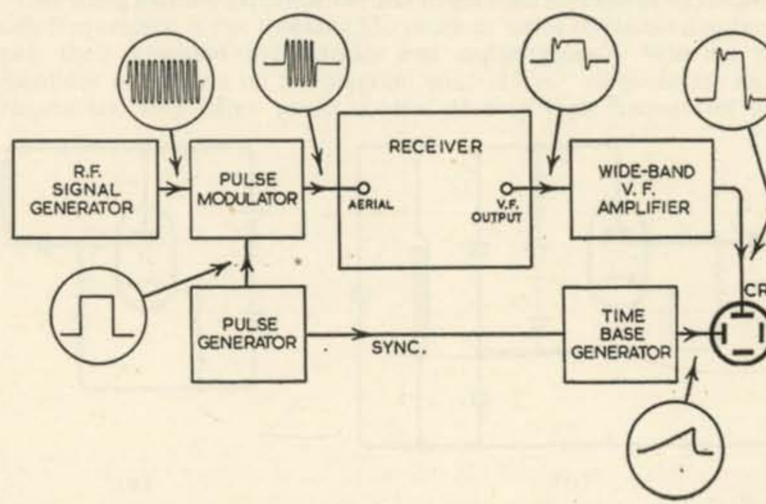


Fig. 11.29—Block diagram of apparatus for overall pulse testing of television receiver

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.

Although complete frequency-response *and phase* data theoretically 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. 11.29 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. 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 of such equipment, and a discussion of results, is given by M. V. Callendar in *Wireless World*, February 1952, pp. 42–45.

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

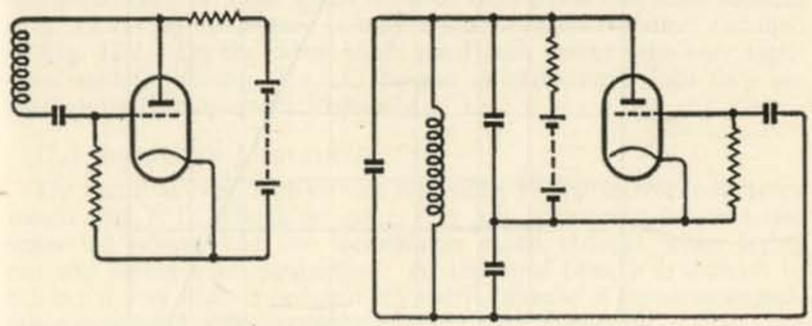


Fig. 12.1—The oscillator circuit which, according to the components actually visible, would be drawn as at a, is better understood when valve interelectrode capacitances are shown (b)

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 primary characteristics 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

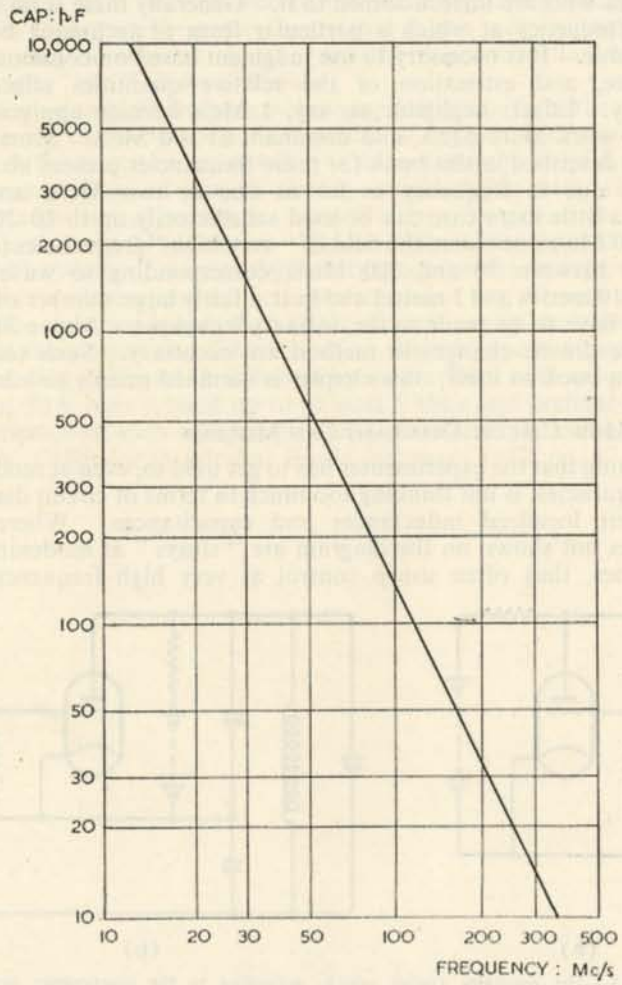


Fig. 12.2—Optimum capacitance of by-pass capacitors, assuming the total length of connections (including the capacitor itself) is one inch

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 in *Wireless World* (29 September 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 about an inch the optimum capacitance is given in Fig. 12.2. This is confirmed in greater detail by R. Davidson in the August 1952 issue of *Wireless World* (pp. 301–4). 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. Thus the short-circuit effected by a capacitor connected as in Fig. 12.3*a* can be improved by connecting as at *b*. And in bush capacitors, which

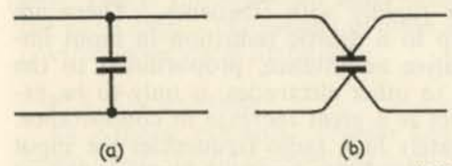


Fig. 12.3—When a by-pass capacitor is connected as at a its impedance is increased by the common connecting leads; this is obviated by connecting as at b

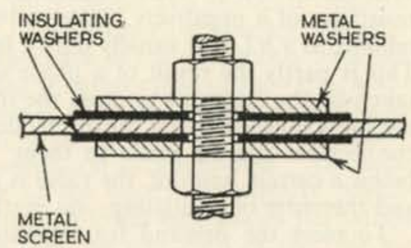


Fig. 12.4—Section of low-inductance by-pass to screen, formed by bushing a lead passing through

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 at v.h.f.) behave as capacitors, 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 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. The performance of valves at v.h.f. and above—especially at 450 Mc/s—is reviewed by E. G. Hamer in *Wireless World*, February 1952, pp. 51–53.

12.5. Noise

Man-made and atmospheric noise declines with increasing frequency, and the limiting factor becomes what is called fluctuation noise

* The properties and applications of transmission lines are presented concisely in *High Frequency Transmission Lines*, by Willis Jackson (Methuen).

(Sec. 14.32), caused by random movement of electrons in circuits and valves. The power of such noise is proportional to absolute temperature (which does not offer much scope for control) and resistance—including the dynamic resistance of tuned circuits. 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. 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 subject of noise and noise factor is now very fully covered in technical literature.

12.6. Oscillators

As a simple example, consider how the use of ordinary valves and components should be modified in an oscillator at a moderate frequency in this band, say about 60 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 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 14 or 16 s.w.g.—shaped on a rod $\frac{5}{8}$ 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 more 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 ½-in tube. Where the resistance 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

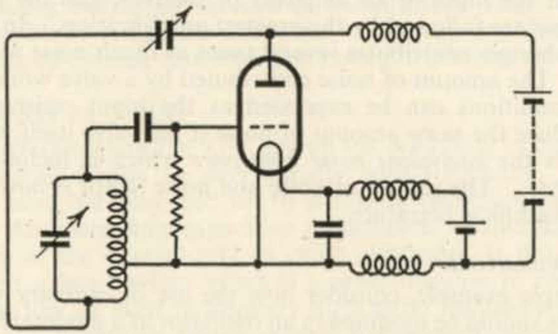


Fig. 12.5—Typical oscillator circuit for v.h.f.

to use a special type of valve. Fig. 12.6 shows the circuit and Fig. 12.7 the layout of a portable sender using an " acorn " valve for 375 Mc/s (80 cm wavelength). The aerial is merely a piece of wire cut to a half-wavelength, which can be excited as shown (current excitation) or by bringing one end near a high-potential part of the coil (voltage excitation). But acorn valves are fragile and expensive to make, and have been largely superseded by other types. The EC32 is a glass-based triode suitable for oscillation up to 400 Mc/s, and there are many types—some comprising a balanced pair of valves for push-pull oscillators —suitable for generating an appreciable 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.

With standard-signal generators the chief difficulty is to get the signal where one wants it and nowhere else. Attenuators of the conventional types

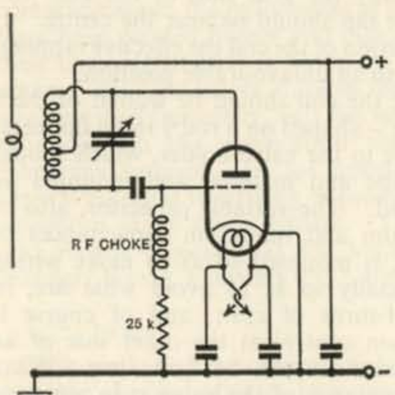


Fig. 12.6—Circuit of sender shown in Fig. 12.7

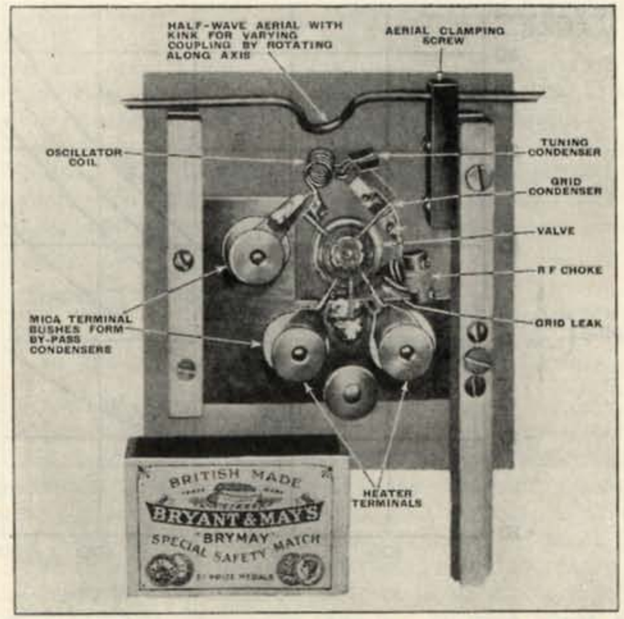


Fig. 12.7—Simple 375-Mc/s sender using an acorn valve. The matchbox indicates the scale. (The dark square is a background, and not part of the apparatus)

become difficult (but see Fig. 4.24) and the waveguide type comes into its own (Sec. 6.22).

12.7. 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 per cent 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. The receiver corresponding to the oscillator shown in Fig. 12.7 is practically the same except for a high-resistance grid leak to induce squegging, or intermittent oscillation, which provides the super-regenerative effect. A review of its theory appeared in *Wireless World*, June 1946, pp. 182–6, and a simple v.h.f. receiver was described in the same journal in January 1947, pp. 15–17.

* 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.*, Pt. III, December 1945, pp. 291–5.

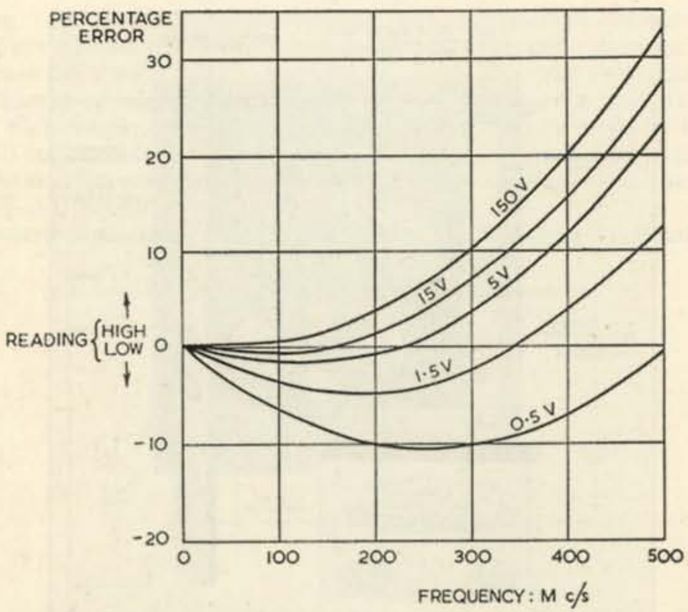


Fig. 12.8—Valve-voltmeter error at v.h.f., due to combined effect of transit time and input resonance, at five indicated r.m.s. voltages. (General Radio type 1800-A)

12.8. 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. 5.12). 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. The behaviour of the diode at v.h.f. was investigated by E. C. S. Megaw*; see also the summary of diode-voltmeter errors in general referred to in Sec. 5.12. 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.

* *Wireless Engineer*, 1936; February, pp. 65–72, March, pp. 135–146, and April, pp. 201–4.

Silicon rectifiers have been used for metering as high as 10,000 Mc/s, but within the limits of the v.h.f. band germanium is more suitable. Compared with the thermionic diode, it has the advantages of small size and no heater, which both facilitate applying it close to the work with the minimum of undesired circuital effects. Its input capacitance is very low, and liability to bring in hum is absent. And there is no "zero" output. On the other hand the back resistance is lower and less consistent (this matters less at v.h.f. because impedances tend to be low generally) and the maximum allowable voltage is less. The higher

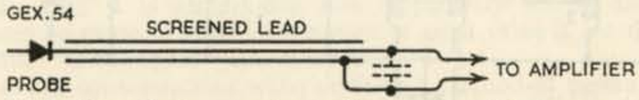


Fig. 12.9—Simple germanium rectifier head for v.h.f. voltmeter

the resistance the lower the voltage, and vice versa. 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.

The applications of germanium rectifiers in the laboratory have yet to be fully explored, but there seems to be considerable scope.

The high-vacuum 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 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.

12.9. 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. 7.9). 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. A somewhat different system, not yet commercially available, which was described by J. E. Houldin on behalf of the G.E.C. Research Laboratories in *Proc. I.E.E.*, Pt. III, November 1952, pp. 389–399, covers the range 1 Ω to 100 kΩ at 50–500 Mc/s by comparison with a standard 100-Ω resistor.

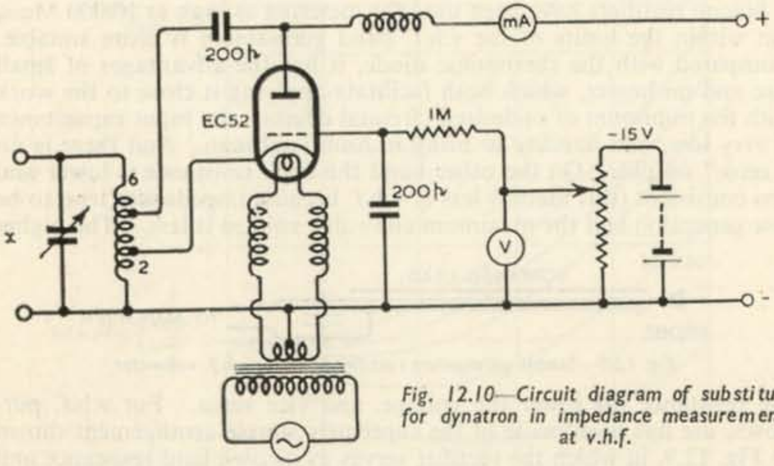


Fig. 12.10—Circuit diagram of substitute for dynatron in impedance measurements at v.h.f.

Measurements at frequencies at which lumped-circuit technique 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. 9.23 to 9.32 can be used, if sufficient care is taken with circuit layout and components. The Hartshorn and Ward apparatus (Sec. 9.26), 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. 9.26, avoid some of the difficulties.

The frequency limit of the dynatron is discussed by G. A. Hay in *Wireless Engineer*, November 1946, pp. 299–305. 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. 4.14 and 4.15 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 the television waveband. The coil consists of six turns of 16 s.w.g. copper, wound as a solenoid ½ 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 measurement of characteristic impedance, 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

$$C = \frac{10^6}{4\pi^2 f^2 L} = \frac{25{,}330}{f^2 L} \qquad [\text{pF};\ \mu\text{H};\ \text{Mc/s}]$$

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

$$C = \frac{1{,}583}{f^2} \qquad [\text{pF};\ \text{Mc/s};\ L = 16\,\mu\text{H}]$$

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. 9.32). 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 Figs. 14.1 and 14.8. 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

* *Radio Data Charts*, by R. T. Beatty and J. McG. Sowerby (Iliffe).

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, but a special vector slide-rule has been devised.†

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.

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 per cent instead of perhaps 1 per cent. Or, having performed a measurement of r.f. resistance by a method that can be

* *Graphical Construction of Vacuum Tube Circuits*, by A. Preisman (McGraw-Hill).

† Blundell Rules, Ltd., Chaul End Lane, Luton, Beds.

depended upon within perhaps 5 per cent, 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, " $10{,}250 \pm 25$ ". 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. 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.*

An advantage of the slide-rule is that it 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 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

* B.S. 1957 : 1953, *The Presentation of Numerical Values* (*B.S.I.*).

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. 9.20, 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, ω = 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).

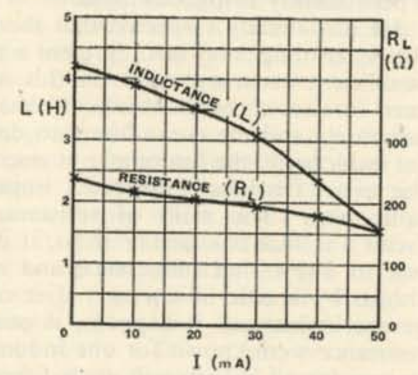


Fig. 13.1—Example of experimental results plotted to aid interpretation

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 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 amps is passed through 100 ohms, 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 amps were passed through 50 ohms; it

is known that it would be exactly twice as much as with 2 amps through 100 ohms. 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 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 very wide range of R—say 0·1 to 100 MΩ. This is useful to know when designing valve voltmeters, even though it is only an empirical law.

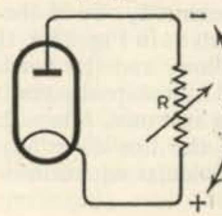


Fig. 13.2—If the electrodes of a diode are joined through a resistance R a voltage V appears across it. The relationship between R and V in a typical sample is plotted in Fig. 13.3

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

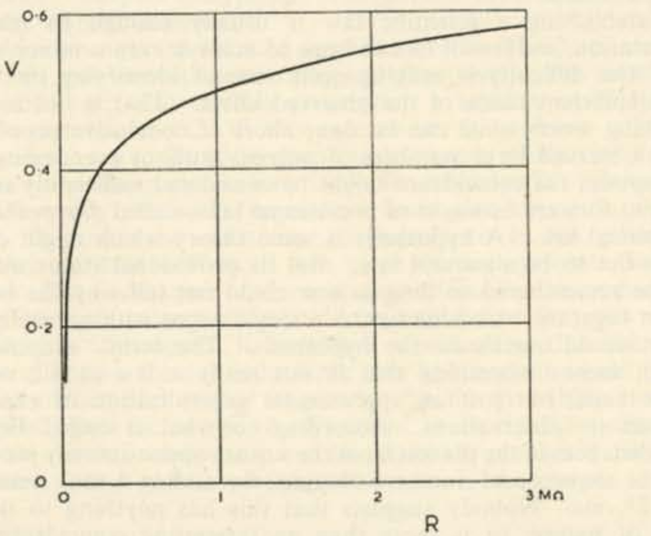
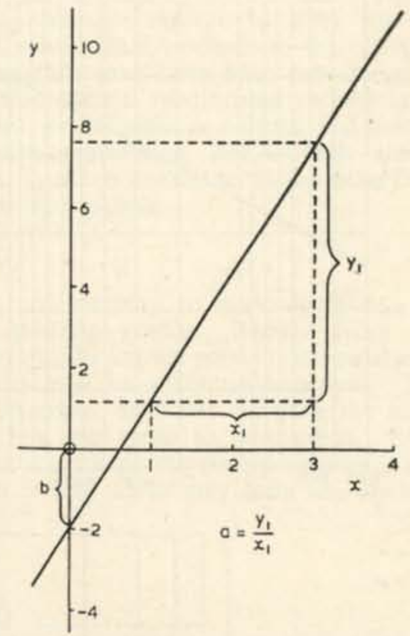


Fig. 13.3—Voltage across a typical diode with normal cathode temperature (Fig. 13.2) plotted against anode-to-cathode resistance, using linear scales

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 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

$$y = 3{\cdot}25x - 2$$

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



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:

$$V = 0{\cdot}16 \log_{10} R + 0{\cdot}50 \qquad [M\Omega]$$

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 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_1 varies as f^{-2}; i.e., is inversely proportional to the square of the frequency. This conclusion can be checked by plotting $1/R_1$ 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—$1/R_1 \propto f^2$—holds good in many different experiments with many

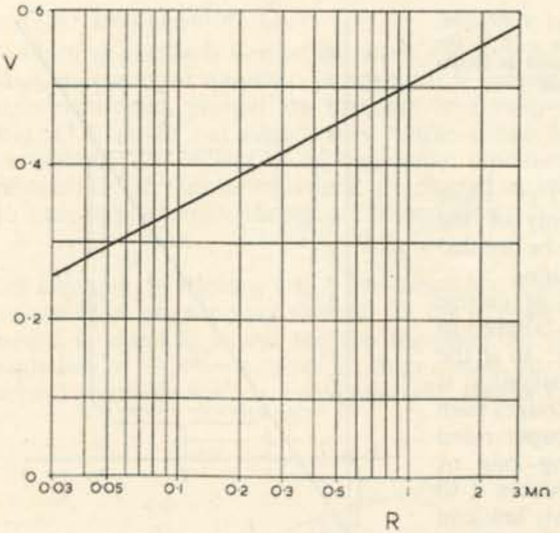


Fig. 13.5—If the curve in Fig. 13.3 is replotted on a logarithmic resistance scale, as in Fig. 13.4, it becomes a very nearly straight line and can be fitted with an equation

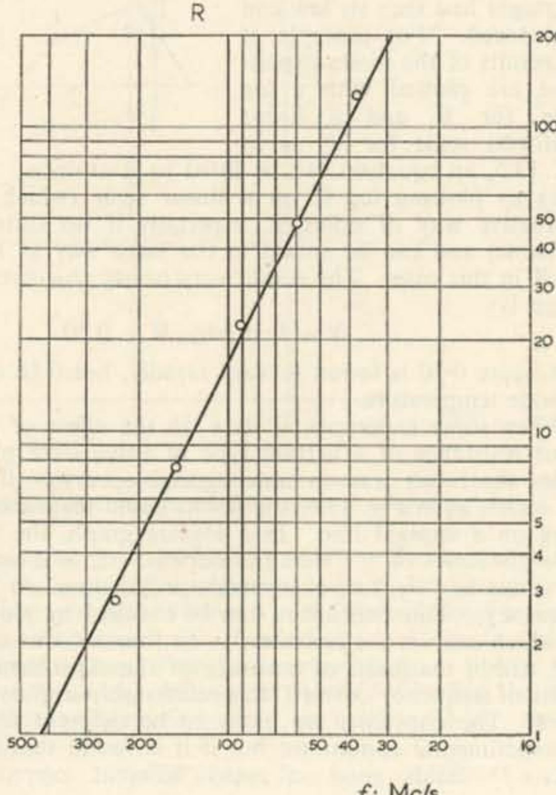


Fig. 13.6—Another example of deriving a law from experimental results; here the input resistance of a valve is plotted against frequency, both with logarithmic scales; and the slope of the line shows that R is inversely proportional to the square of the frequency

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 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

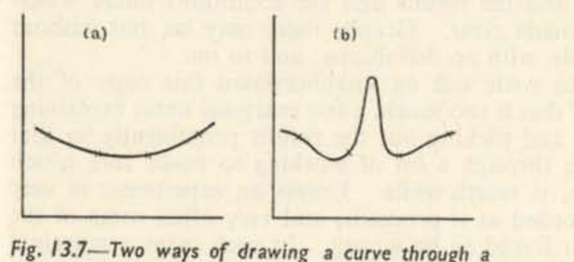


Fig. 13.7—Two ways of drawing a curve through a series of plotted points. Which is right?

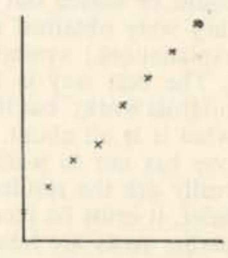


Fig. 13.8—If it were not for one point, a straight line could be drawn through them all. Is the odd reading an error, or does it point out a fact?

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."

* *Electrical Measurements*, p. 8 (Chapman & Hall, 1952).

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.

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 is becoming 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 *Wireless Engineer*, and especially the indexes to them published annually.

CHAPTER 14

For Reference

14.1. Units

For a long time students of all branches of electrical engineering have been 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 κ_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 κ_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 κ for various materials, based on 1 for empty space, as relative or specific quantities, to be multiplied by μ_0 and κ_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.

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.*, Pt. I, September 1950, pp. 235–272.

14.2. Symbols, Abbreviations and Unit Equivalents

As far as possible the symbols used in this book are British Standard.*

* B.S. 560: *Engineering Symbols and Abbreviations* (B.S.I.).

The list that follows includes most of those in common use. It is important to observe the 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 latter is correctly used only after a number specifying a particular amount, as 160 μH (standing for 160 microhenries). If the same thing were used in an equation it would mean "160 × permeability × magnetizing force".

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	
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)

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 $\hat{\imath}$)	peak value
I_{p-p}	peak-to-peak value
I_{av}	mean value
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	κ	$\kappa_0\kappa_s$	same as κ_s
Electric space constant	κ_0	$1/(36\pi \times 10^9)$ approx.	1
Relative or specific permittivity	κ_s	same as κ in c.g.s.	same as κ
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:*
M	mega-	10^6
k	kilo-	10^3
m	milli-	10^{-3}
μ	micro-	10^{-6}
*n	nano-	10^{-9}
p (or $\mu\mu$)	pico-	10^{-12}

* This submultiple is not yet universally used, but is convenient for enabling capacitor values to be specified without decimal fractions.

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)	= 0·9144 metre = 3 feet = 36 inches
			1 foot (ft)	= 30·48 cm = 12 inches
			1 inch (in)	= 2·540 cm = 1,000 mils
	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 inches	1 cubic inch (cu in or in^3)	= 16·39 cm^3

Table 14.4 (*continued*)

Angle	1 radian	$= \frac{360}{2\pi}$ or 57·29 degrees	1 revolution	$= 2\pi$ 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		

* *at sea level at 45° latitude*

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		
Power	1 kilowatt (kW)	= 1·341 horsepower = 737·6 ft lb/sec	1 horsepower (hp)	= 0·7457 kW = 550 ft lb/sec

Temperature

Table 14.5

	Degrees:		
	Kelvin (°K)	Centigrade (°C)	Fahrenheit (°F)
Absolute zero	0	− 273	− 514
Freezing water	273	0	32
Boiling water	373	100	212
At normal pressure			

$°K = °C + 273$ $\qquad$ $°F = \frac{9}{5}°C + 32$

Light See Sec. 14.30.

Greek Alphabet

Α	α	alpha	Η	η	eta	Ν	ν	nu	Τ	τ	tau
Β	β	beta	Θ	θ	theta	Ξ	ξ	xi	Υ	υ	upsilon
Γ	γ	gamma	Ι	ι	iota	Ο	o	omicron	Φ	ϕ	phi
Δ	δ	delta	Κ	κ	kappa	Π	π	pi	Χ	χ	chi
Ε	ϵ	epsilon	Λ	λ	lambda	Ρ	ρ	rho	Ψ	ψ	psi
Ζ	ζ	zeta	Μ	μ	mu	Σ	σ or s	sigma	Ω	ω	omega

Meanings of some Greek letters commonly used:

- α overall voltage gain, especially with reference to negative feedback loops*
- β fraction of α fed back
- δ (1) loss angle; (2) logarithmic decrement ($\simeq \pi/Q$).
- δ, $\triangle$ } a small difference or increment
- θ an angle
- κ permittivity
- λ wavelength
- μ (1) permeability; (2) valve amplification factor; (3) one millionth of
- π circumference/diameter of circle (= 3·14159 . . .).
- Π product of terms such as . . .
- ρ resistivity
- Σ sum of terms such as . . .
- τ time period
- Φ magnetic flux
- ϕ phase angle
- Ω ohm
- ω $2\pi f$

* Some authorities who ought to know better use μ for this purpose—as if that symbol were not already overworked! This usage is particularly confusing because μ is the world standard symbol for valve amplification factor.

Mathematical Symbols and Abbreviations (see Sec. 14.33)

- $=$ equal to
- $\equiv$ identically equal to (i.e., regardless of varying values)
- $\simeq$ approximately equal to
- $<$ less than
- $<<$ much less than
- $\nless$ not less than
- $\leq$ equal to or less than
- $>$ greater than; etc.
- $\propto$ proportional to
- ∞ infinity
- $|\ |$ magnitude or numerical value of term enclosed (i.e., regardless of sign or vectorial direction)
- e base of natural logarithms (= 2·71828 . . .).
- 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

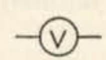
Note: $10^3 = 1{,}000$ $\quad 10^0 = 1$ $\quad 10^{\frac{1}{2}} = \sqrt{10}$
$10^2 = 100$ $\quad 10^{-1} = 0{\cdot}1$ $\quad 10^{-\frac{1}{2}} = 1/\sqrt{10}$
$10^1 = 10$ $\quad 10^{-2} = 0{\cdot}01$ $\quad 10^{\frac{3}{2}} = \sqrt{10^3}$
etc.

- $f(x)$ or $F(x)$ function of x
- $f'(x)$ or $\frac{d}{dx}f(x)$ or $D_x f(x)$ differential coefficient, or slope, of $f(x)$
- $\int f(x)dx$ integral of $f(x)$
- $\sin^{-1}x$ angle whose sine is x
- Π product (of all terms of type specified)
- Σ sum (of all terms of type specified)

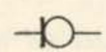
Graphical Symbols

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


piezo crystal

 electrolytic capacitor (white plate positive)

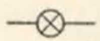
 cell (long stroke positive)

 black spot denotes tube is gas-filled

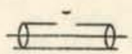
 thermistor

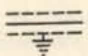
 transistor

 connection to chassis

 indicator lamp

 dust core

 coaxial cable

 or  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.

Miscellaneous Symbols

A	voltage gain or amplification (alternative to α)
B	feedback factor (alternative to β)
c	speed of light, $= 299{,}790$ km/s
f_o	frequency of oscillation
f_r	frequency of resonance
Δf	frequency difference
f'	frequency off-tune
g	acceleration due to gravity (at earth's surface), $= 9{\cdot}81$ metres/sec^2
h_e	effective height of aerial
k	coefficient of coupling
K	modulation depth
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)
ϕ	phase angle (between vectors representing alternating current and voltage)
$\cos\phi$	power factor, R/Z, $= \sin\delta$
δ	loss angle, $= 90° - \phi$
$\tan\delta$	dissipation factor or loss tangent, $= \cot\phi$, $=$ R/X, $= 1/Q$

When $\delta <$ about 10°, $\tan\delta \simeq \sin\delta \simeq \delta$ (in radians)

Valve Abbreviations

The following are the most important conventions set out fully in B.S. 1409: *Letter Symbols for Electronic Valves* (B.S.I.):

k	cathode
g	grid
g_1	first grid (nearest cathode)
g_2	second grid; and so on
a	anode
μ	amplification factor
r_a	anode a.c. resistance or slope resistance
g_m	mutual conductance
g_c	conversion conductance (of frequency changer)

Symbols are frequently combined, thus:

I_a	Anode current
V_{g2}	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.

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
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
h.t.	high tension
i.f.	intermediate frequency
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.	vision (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 official nomenclature for *radio* frequencies:

Table 14.6

Description	*Abbreviation*	*Frequencies Included*	*Corresponding Wavelengths*
Very low	v.l.f.	below 30 kc/s	over 10,000 metres
Low	l.f.	30–300 kc/s	1,000–10,000 metres
Medium	m.f.	300–3,000 kc/s	100–1,000 metres
High	h.f.	3–30 Mc/s	10–100 metres
Very high	v.h.f.	30–300 Mc/s	1–10 metres
Ultra-high	u.h.f.	300–3,000 Mc/s	10–100 cm
Super-high	s.h.f.	3,000–30,000 Mc/s	1–10 cm

The description " extra-high frequency " (e.h.f.) is advocated by *Wireless World* for all above 30 Mc/s.

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 ".

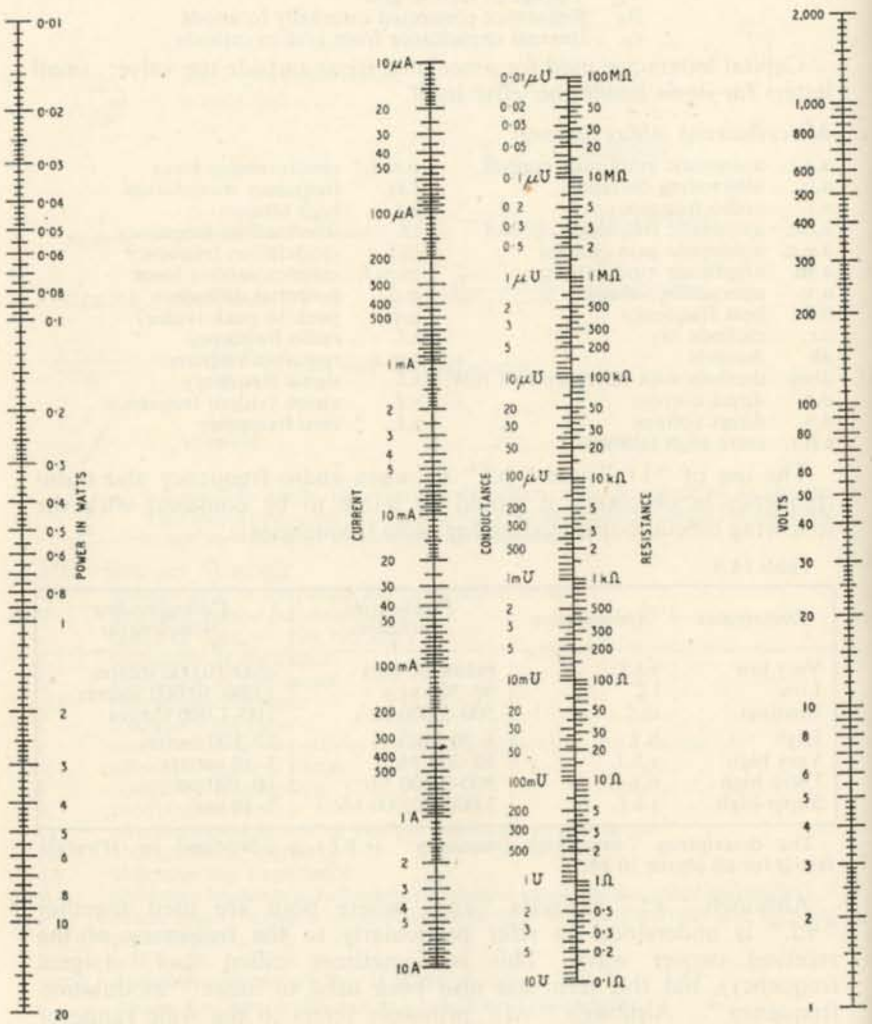


Fig. 14.1—This chart (adapted from No. 42 in Radio Data Charts, *5th ed.) is used by connecting any two known quantities by a ruler or a stretched thread, whereupon the intersections indicate the other three quantities. E.g., a 10-kΩ resistor rated at ½ W is seen to be limited to 7 mA and 70 V. Its conductance is also shown as 100 µ℧ (℧ being the proposed symbol for "mho")*

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 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 not strictly correct it is convenient.

An alternative form (Sec. 14.13) is

$$I = EG$$

This is convenient for parallel circuits, for the total conductance, G, is the sum of the separate conductances in parallel. For relating E, I, R, G and P, see Fig. 14.1.

Ref: " Ohm's Law ", *Wireless World*, August 1953, pp. 383–6.

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. 14.2) one can write down Kirchhoff equations, such as:

$$I_s - I_1 - I_3 = 0$$
$$I_1R_1 + I_dR_d - I_3R_3 = 0$$

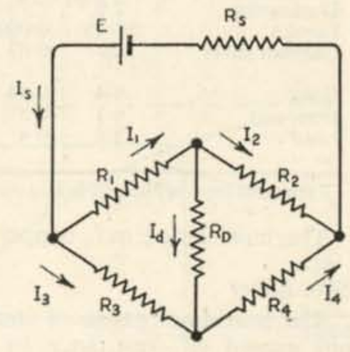


Fig. 14.2

The three loops and four junctions give seven simultaneous equations for solving the six unknown currents.

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 1·72 $\mu\Omega$-cm, = 0·0172 $\mu\Omega$-metre, increasing 0·4 per cent 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	0·0018	Manganin	27	0·002
Aluminium, hard-drawn	1·63	0·39	Mercury	56	0·09
Brass	4·4	0·15	Nichrome	58	0·04
Concordin	60	0·17	Nickel	6·3	0·43
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	0·1
Duralumin	2·0	—	Platinoid	20	0·03
Eureka	28–29	0·0008	Platinum	6·3	0·36
German silver	20	0·03	Silver	0·94	0·38
Gold	1·4	0·34	Tin	6·7	0·45
Iron, soft	6·1	0·5	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 z.f. resistance by more than 1 per cent 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:

$$C = \frac{A\kappa}{t} \qquad \begin{array}{l} C \text{ in farads} \\ A \text{ in sq metres} \\ t \text{ in metres} \end{array}$$

$\kappa = \kappa_0\,\kappa_s$, where $\kappa_0 = 10^7/4\pi c^2 = 8{\cdot}854 \times 10^{-12}$ 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:

$$C = \frac{0{\cdot}0885\, A\, \kappa_s}{t} \text{ picofarads}$$

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 κ_s, the specific permittivity, or dielectric constant, are given in Table 14.9.

Capacitance where more than two conductors are involved—see Sec. 9.15.

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 of C after t seconds will be

$$V_t = E - (E - V_0)e^{-t/RC}$$

(e = base of natural logarithms = 2·718 . . .)

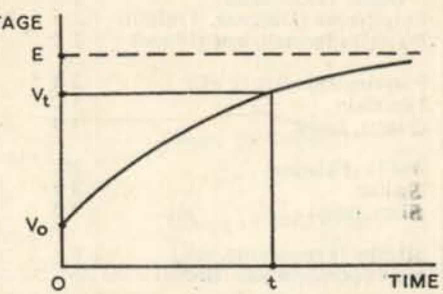


Fig. 14.3

Alternatively $\quad 2{\cdot}30 \log_{10} \dfrac{E - V_t}{E - V_0} = -\dfrac{t}{RC}$

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 per cent $(=1-1/e)$ 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

κ_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*)	κ_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	10^8	
Glass, plate	6–7	40	10^6	10^7
Methyl methacrylate (Perspex)	2·6–3·4	60	10^8	10^9
		600	50	
Mica	4·5–7	2–3	10^8	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	2–5		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 κ_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.
$E = 4{\cdot}44\ \Phi_{max} N f$	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
	l in cm
$B = \mu_s H$	B in gauss
$\Phi = BA$	Φ in maxwells (" lines ")
$L = \dfrac{\Phi N}{I \times 10^8} = \dfrac{1{\cdot}257\ \mu\ AN^2}{l \times 10^8}$	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, $\dfrac{N}{E} = \dfrac{1}{4{\cdot}44\ \Phi_{max} f}$ $\qquad$ (Φ_{max} in webers)

or $\dfrac{10^8}{4{\cdot}44\ \Phi_{max} f}$ $\qquad$ (Φ_{max} in maxwells)

Impedance Z_s in secondary circuit of ideal transformer is equivalent to $Z_s (N_p/N_s)^2$ in the primary circuit, where N_p and N_s are primary and secondary turns respectively.

Appropriate transformer ratio, $n = \dfrac{N_P}{N_S} = \sqrt{\dfrac{Z_P}{Z_L}}$

where Z_L = load impedance

Z_P = desired impedance from primary side

Leakage inductance $= (1 - k^2)\,L_P$

where k = coefficient of coupling: see Sec. 14.19.

For equivalents of a.f. transformers in valve circuits, see Sec. 14.16.

14.10. Inductance

The universal curve sheet, Fig. 14.4, 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

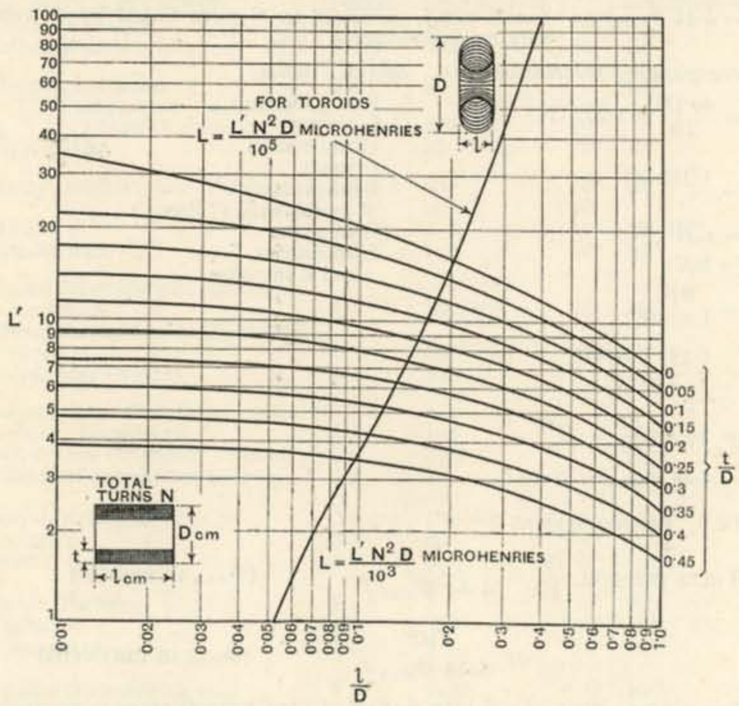


Fig. 14.4—Chart compiled from standard formulae by F. Charman, by which the self-inductance of coils of most shapes can be easily found

the correct number of turns of wire on a former of given shape for a required inductance to be calculated. The tables in Sec. 14.39 then enable a suitable gauge of wire to be selected.

The following formula* for single-layer coils, which is accurate to within 1 per cent when $l/D >$ about 0·4, fills in where the $t/D = 0$ curve leaves off, at $l/D = 1$:

$$L = \frac{N^2 D_{cm}}{101(l/D + 0{\cdot}45)} = \frac{N^2 D_{in}}{40(l/D + 0{\cdot}45)}$$

D_{cm} = dia. in cm.

D_{in} = dia. in in.

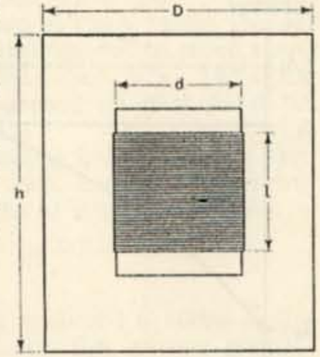


Fig. 14.5

Coil Screening

The inductance of a coil screened by a non-magnetic cylindrical can is given approximately by multiplying the unscreened inductance by the following factor,† in which the dimensions are as shown in Fig. 14.5.

$$\frac{D^3 - d^3}{D^3}\left[1 - \left(\frac{l}{2h}\right)^2\right]$$

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 $M = \tfrac{1}{2}(L - L_1 - L_2)$

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 continuous between those points; and similarly for the others.

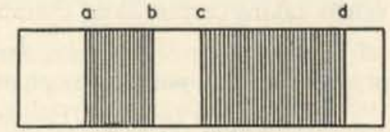


Fig. 14.6

For more comprehensive data for calculation of L, M, and C, refer to F. Langford-Smith's *Radio Designer's Handbook* (Iliffe).

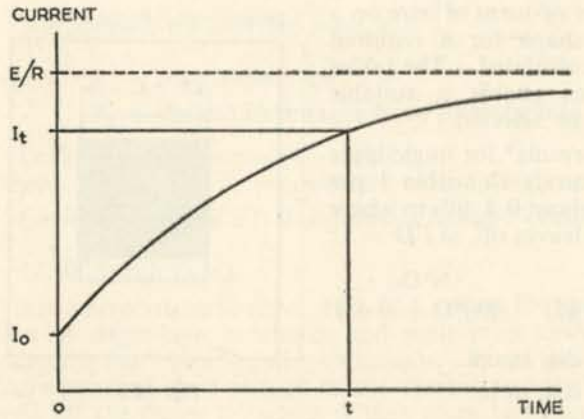
Inductive Transients

If L henries, carrying an initial current of I_0 amperes, is connected to

* Due to H. A. Wheeler, *Proc. I.R.E.*, October 1928, p. 1398.

† Due to W. G. Hayman, *Wireless Engineer*, April 1934, p. 189.

Fig. 14.7



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 $\qquad 2{\cdot}30 \log_{10} \dfrac{E/R - I_t}{E/R - I_0} = -\dfrac{Rt}{L}$

The quantity L/R is the *time constant*, equal to the time occupied by the first 63·2 per cent ($=1-1/e$) 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. 359).

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: $\sin (\omega t + \pi/2) = \cos \omega t$)

$$I = \frac{I_{max}}{\sqrt{2}} = 0{\cdot}707\ldots\ I_{max}$$

$$\text{So} \quad I_{max} = 1{\cdot}414\ldots\ I$$

$$\text{and} \quad I_{p-p} = 2{\cdot}828\ldots\ I$$

$$I_{av} = \frac{2}{\pi} I_{max} = 0{\cdot}637\ldots\ I_{max}$$

$$\text{So} \quad I = \frac{\pi}{2\sqrt{2}} I_{av} = 1{\cdot}11\ldots\ I_{av}$$

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. 5.11).

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 ", *Wireless World*, October 1946, pp. 335–8.

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, but not of r.m.s. values nor of peak values (unless the peaks all coincide). 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 ", *Wireless World*, March 1952, pp. 117–120.

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.

$$\text{So} \quad X = X_L + X_C$$

$$\text{where } X_L = \omega L \text{ and } X_C = -1/\omega C \text{ (see Fig. 14.8)}$$

$$\text{and} \quad \omega = 2\pi f$$

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.

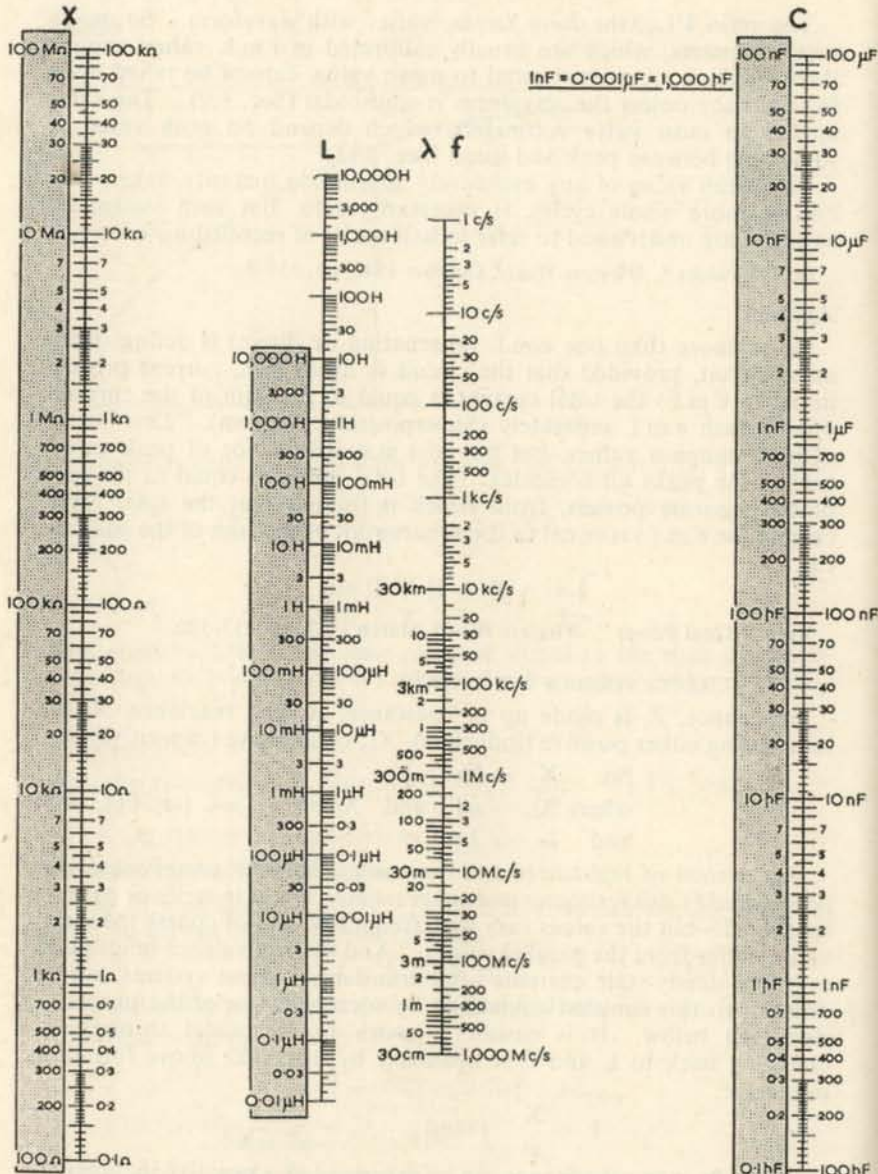


Fig. 14.8—This chart, used in the same way as Fig. 14.1, shows the reactance of a given inductance or capacitance at a given frequency, or the inductance and capacitance required to tune to a given frequency or wavelength, or vice versa. Use either all the shaded scales or none of them. E.g: Find the 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 Ω; connecting this with 500 kc/s shows L and C to be 160 μH and 640 pF respectively

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:*

$$G = 1/R \quad \text{and} \quad B = -1/X$$
$$B = B_C + B_L$$
$$\text{where } B_C = \omega C \quad \text{and} \quad B_L = -1/\omega L$$
$$Y = 1/Z = \sqrt{G^2 + B^2}$$

To translate parallel circuit elements into their series equivalents, and vice versa, these relationships are very useful:

$$R_s = X_p \frac{R_p X_p}{R_p^2 + X_p^2} \qquad X_s = R_p \frac{R_p X_p}{R_p^2 + X_p^2}$$
$$R_p = \frac{R_s^2 + X_s^2}{R_s} \qquad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

E.g: Reduce the circuit shown in Fig. 14.9 to its simplest equivalent at 400 c/s. The resistance of the coil is represented by the 360 Ω, and the resistances of the capacitors are assumed to be negligible.

The equivalent reactances are worked out from $X_L = \omega L$, $X_C = -1/\omega C$ (Fig. 14.8). $\omega = 2{,}500$ very nearly. This immediately reduces to Fig. 14.10.

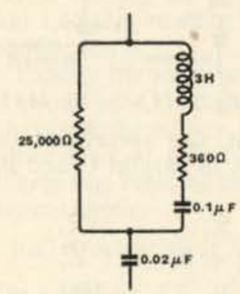


Fig. 14.9

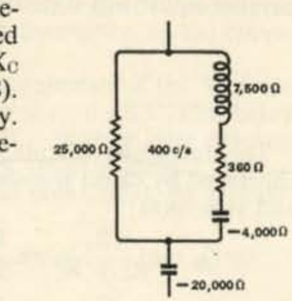


Fig. 14.10

* "Admittance", *Wireless World*, January 1949, pp. 29–31.

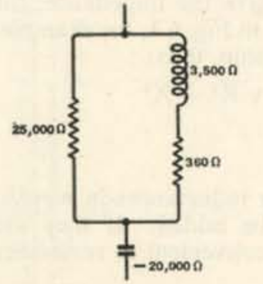


Fig. 14.11

Converting the series elements into parallel equivalents:

$$R_p = \frac{360^2 + 3{,}500^2}{360} = 34{,}500\ \Omega$$

$$X_p = \frac{360^2 + 3{,}500^2}{3{,}500} = 3{,}550\ \Omega$$

Combining the resistances in Fig. 14.12 by adding the reciprocals:

$$R = \frac{25{,}000 \times 34{,}500}{25{,}000 + 34{,}500}$$
$$= 14{,}500\ \Omega \text{ (Fig. 14.13)}$$

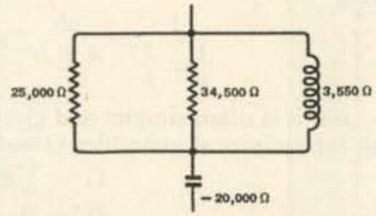


Fig. 14.12

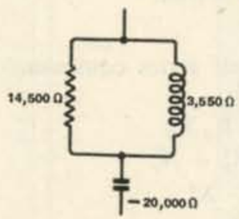


Fig. 14.13

Converting the parallel elements into series equivalents:

$$R_s = \frac{14{,}500 \times 3{,}550^2}{14{,}500^2 + 3{,}550^2} = 820\ \Omega$$

$$X_s = \frac{14{,}500^2 \times 3{,}550}{14{,}500^2 + 3{,}550^2} = 3{,}350\ \Omega$$

Combining reactances in Fig. 14.14 gives the simplest series equivalent (Fig. 14.15); and this can easily be changed into the parallel equivalent if desired.

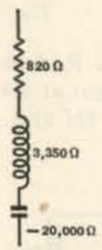


Fig. 14.14

Fig. 14.15

The foregoing calculation, and others like it, can sometimes be shortened by direct transfers from series R and X to parallel G and B, and vice versa:

$$G_p = \frac{R_s}{R_s^2 + X_s^2} = \frac{R_s}{Z_s^2} \qquad B_p = -\frac{X_s}{R_s^2 + X_s^2} = -\frac{X_s}{Z_s^2}$$

$$R_s = \frac{G_p}{G_p^2 + B_p^2} = \frac{G_p}{Y_p^2} \qquad X_s = -\frac{B_p}{G_p^2 + B_p^2} = -\frac{B_p}{Y_p^2}$$

The 25 kΩ, being a parallel element, would be expressed as a

conductance, 40 micromhos. The R_s and X_s in Fig. 14.11 would be transformed straight into G_p and B_p. This G_p would simply be added to the 40 $\mu\mho$ to give the total conductance, then it and B_p would be transformed into R_s and X_s as in Fig. 14.14.

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:

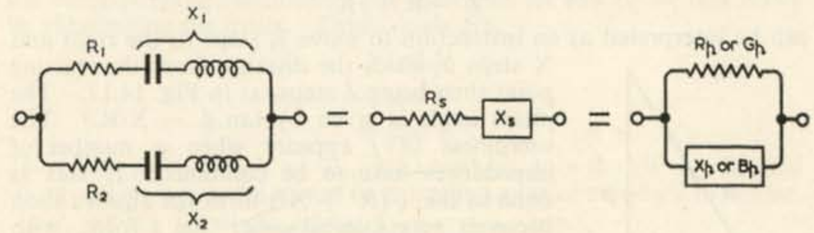


Fig. 14.16

$$R_s = \frac{R_1R_2(R_1+R_2)+R_1X_2^2+R_2X_1^2}{(R_1+R_2)^2+(X_1+X_2)^2} \qquad X_s = \frac{X_1X_2(X_1+X_2)+R_1^2X_2+R_2^2X_1}{(R_1+R_2)^2+(X_1+X_2)^2}$$

$$R_p = \frac{1}{G_p} = \frac{(R_1^2+X_1^2)(R_2^2+X_2^2)}{R_1(R_2^2+X_2^2)+R_2(R_1^2+X_1^2)}$$

$$X_p = -\frac{1}{B_p} = \frac{(R_1^2+X_1^2)(R_2^2+X_2^2)}{X_1(R_2^2+X_2^2)+X_2(R_1^2+X_1^2)}$$

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.

Ref: *Wireless World*, January 1950, pp. 19–20, and February 1950, p. 76.

The j Method

The example given on p. 377 was in numerical values, but if it had been necessary to calculate the impedance of that circuit—or any other except the simplest kinds—in general, using letter symbols, it 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 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. 14.17. 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 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

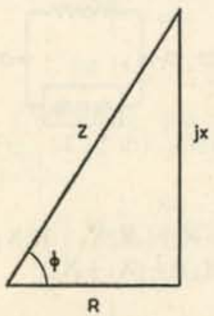


Fig. 14.17

$$(p + jq) + (r + js) = (p + r) + j(q + s)$$

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". *Wireless World*, February 1953, pp. 79–82.

14.13. Duals

It may have been noticed that of the two pairs of equations at the foot of p. 378 each pair is exactly the same as the other, except that the

* E.g., *Basic Mathematics for Radio Students*, by F. M. Colebrook (Iliffe); and *A.C. Network Analysis by Symbolic Algebra*, by W. H. Miller (Classifax Publications). See also "j", *Wireless World*, February 1948, pp. 68–71.

† It is necessary, of course, to have regard for the sign of X, and possibly of R.

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:

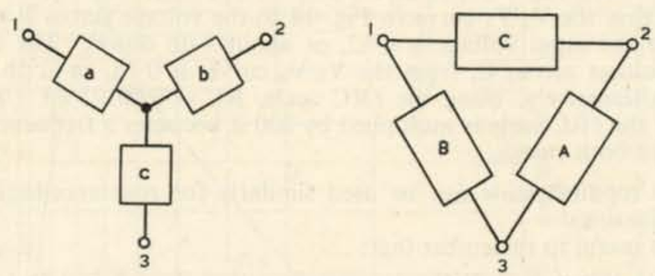
I and V
Z and Y
C and L

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". *Wireless World*, April 1952, pp. 152–5.

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



(a) (b)

Fig. 14.18—(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*
$a = \dfrac{BC}{A+B+C}$	$A = \dfrac{ab+bc+ca}{a}$
$b = \dfrac{CA}{A+B+C}$	$B = \dfrac{ab+bc+ca}{b}$
$c = \dfrac{AB}{A+B+C}$	$C = \dfrac{ab+bc+ca}{c}$

14.15. Standard Response Curves

One of the commonest circuit calculations is finding (for example) the proportion of the input voltage reaching the valve in Fig. 14.19.

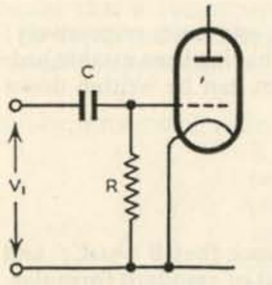


Fig. 14.19—Example of resistance-reactance potential divider

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. Compared with a wholly resistive or wholly reactive divider, it has two complications: X_C varies with frequency, and Z is *not* equal to R + X. 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 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. 14.20, a single curve suffices for all combinations of resistance and one reactance.

E.g: $R = 0{\cdot}25\ M\Omega$, $C = 5\ nF$ ($= 0{\cdot}005\ \mu F$); find loss at 100 c/s.

From Fig. 14.8, X at 100 c/s $= 0{\cdot}318\ M\Omega$. So $R/X = 0{\cdot}785$. From the V_3/V_1 curve in Fig. 14.20 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 for any given case.

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

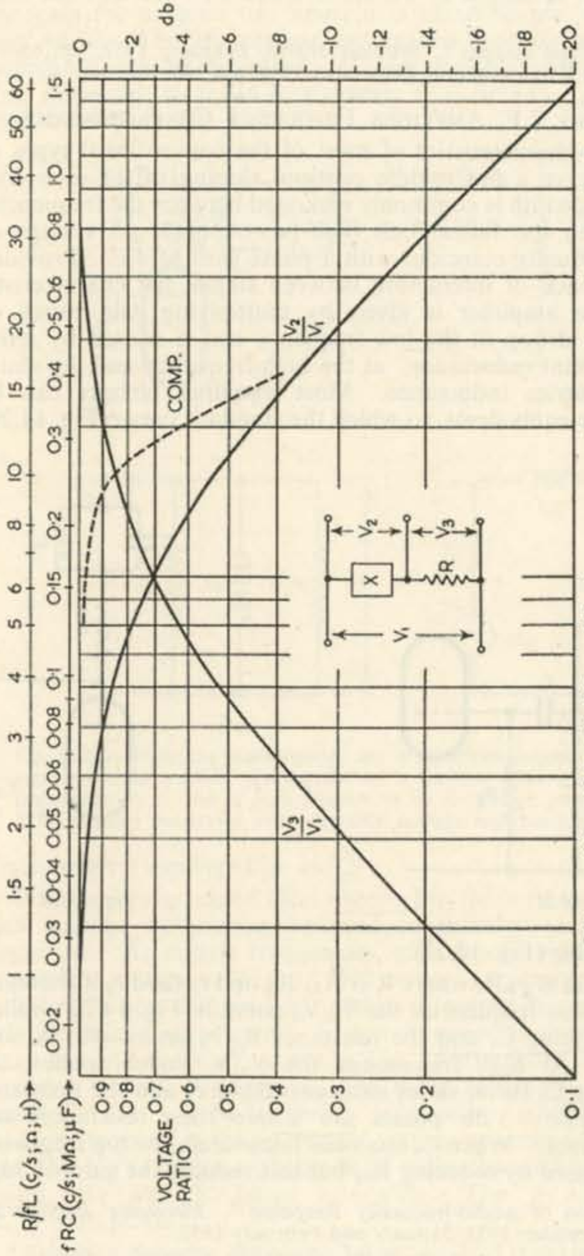


Fig. 14.20 Universal curves whereby the division of voltage between a resistance and a reactance can be read off. This is particularly useful in estimating the performance, at different frequencies, of valve amplifier circuits. The dotted curve refers to the inductance-compensated circuit (Fig. 14.22)

circuits is treated at length by N. H. Crowhurst;* see also the next section.

Refs: "Generalized Graphs", *Wireless World*, February 1952, pp. 65–70. "Equalization", *Wireless World*, September 1949, pp. 348–353.

14.16. A.F. and V.F. Amplifier Frequency Characteristics

The frequency characteristics of most of the conventional types of amplifier stage have a flat middle portion, sloping off at each end. The effective bandwidth is commonly reckoned between the frequencies at which the gain has fallen 3 db (half-power, or $1/\sqrt{2}$ voltage or current). This usually coincides with a phase shift of 45°. Provided there is no feedback or interaction between stages, the characteristic for a multi-stage amplifier is given by multiplying gain ratios or adding db. The droop at the low-frequency end is caused by series capacitance or shunt inductance; at the high-frequency end, by shunt capacitance or series inductance. Most amplifier circuits can be reduced to simple equivalents, to which the standard curve (Fig. 14.20) can be fitted.

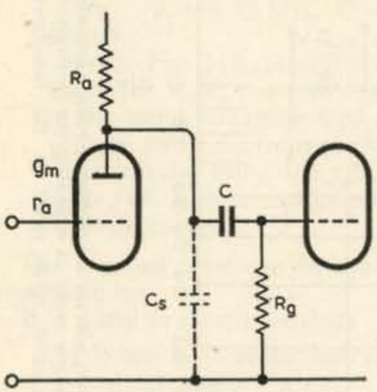


Fig. 14.21

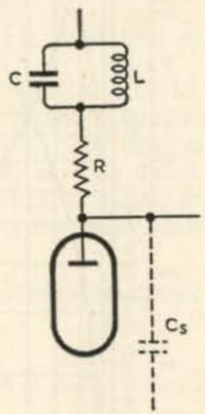


Fig. 14.22

Resistance Coupling (Fig. 14.21)

The "flat" gain is $g_m R$, where R is R_a, R_g and r_a (and r_g if finite) all in parallel. At low frequencies the V_3/V_1 curve in Fig. 14.20 applies, the capacitance being C, and the resistance R_g in series with R_a and r_a in parallel.† At high frequencies the V_2/V_1 curve applies, the capacitance being C_s (total shunt stray capacitance) and the resistance R as above. The -3 db points are where these resistances and reactances are equal. When C_s has been minimized, the top frequency can be further raised by reducing R_a, but this reduces the gain, so high

* "The Prediction of Audio-frequency Response", *Electronic Engineering*, November and December 1951, January and February 1952.

† *Radio Engineering*, 3rd ed., by F. E. Terman (McGraw-Hill), p. 179.

g_m is needed. To raise the top limit for v.f. amplification (or to increase the gain for a given top limit) it is usual to use inductance, as in Fig. 14.22. The appropriate inductance depends on whether or not overshoot can be tolerated. If not, the best value (for maximum flat response) is $0{\cdot}414\,R^2 C_s$ with $C = 0{\cdot}35\,C_s$, giving the dotted characteristic in Fig. 14.20.*

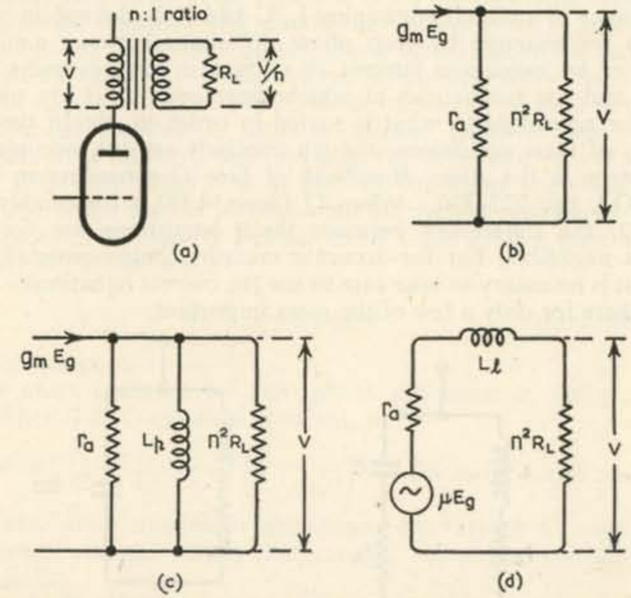


Fig. 14.23—Neglecting capacitance, the transformer-coupled stage (a) can be represented at medium frequencies by b (current generator circuit), at low frequencies by c, and at high frequencies by d (voltage generator circuit). L_p is the primary inductance and L_l the total leakage inductance referred to primary

Transformer Coupling (Fig. 14.23)

Transformer assumed ideal except for finite primary inductance L_p and leakage inductance (referred to primary) L_l. Capacitance is neglected. At middle frequencies, gain can be reckoned from either equivalent voltage or (Fig. 14.23*b*) current generator; but transformer coupling is now seldom used except in output stages, where only the relative gain is significant. At low frequencies the equivalent circuit is as at *c* and the V_3/V_1 curve in Fig. 14.20 applies if the R/X scale is changed to X/R, X being ωL_p and R the resistance of r_a and n^2R_L in parallel. The -3 db frequency is therefore $r_a n^2 R_L/2\pi L_p(r_a+n^2R_L)$. At high frequencies the equivalent voltage generator (*d*) is appropriate, and the V_2/V_1 curve with the R/X scale changed to X/R applies, X being ωL_l and R being $r_a + n^2R_L$. The -3 db frequency is $(r_a + n^2R_L)/2\pi L_l$.

* *Television Receiving Equipment*, 3rd ed., by W. T. Cocking (Iliffe), Ch. 9.

With step-*up* transformers L_l tends to resonate with secondary stray capacitance to give a peak before cut-off. Curves for this and other conditions are given in the excellent treatment of the subject by F. E. Terman referred to above.

14.17. 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. 14.18) is reasonably large, say >10, the differences between these conditions are for most purposes negligible, but for accurate measurements—especially with 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.

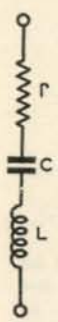


Fig. 14.24

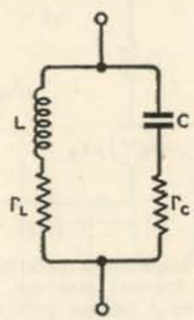


Fig. 14.25

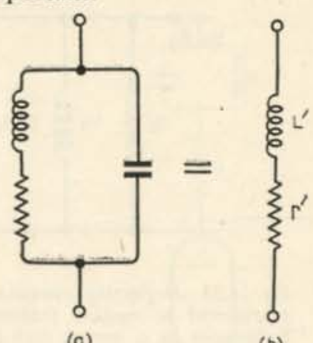


Fig. 14.26

Series Resonance (Fig. 14.24) 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:

$$f_r = 159{\cdot}2/\sqrt{LC} \qquad \text{[kc/s; μH; μF] or [Mc/s; μH; pF]}$$

Or 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. 14.25) 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.

Fig. 14.26 illustrates a useful equivalent. At any frequency f, a (resonating at f_r) is equivalent to b if

$$L' = \frac{L}{1 - (f/f_r)^2} \quad \text{and} \quad r' = \frac{r}{[1 - (f/f_r)^2]^2}$$

If either arm of a parallel tuned circuit is tapped (Fig. 14.27), the

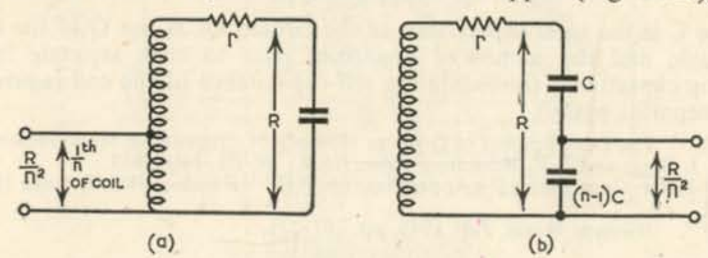


Fig. 14.27

impedance across the tapped-off portion at resonance is still resistive and is proportional to the square of the tapping ratio.

Ref: "Resonance Curves", *Wireless World*, January 1953, pp. 29–33.

14.18. 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 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. 14.28) are shunted by self-capacitance C_0, as in the type of Q meter shown in Fig. 7.18, 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. 9.26) 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.*, Pt. III, July 1949.
“ Q-meter Controversy ” (Appendix), by P.H. *Wireless World*, June 1949, p. 217.
“ Q ”. *Wireless World*, July 1949, pp. 267–271.

14.19. Response 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

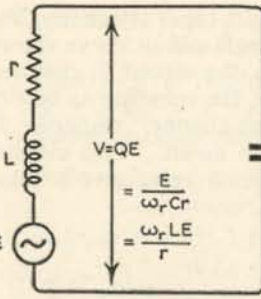


Fig. 14.28

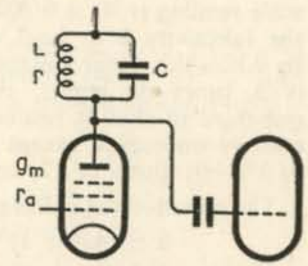


Fig. 14.29

same as those in Fig. 14.20. 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. 14.29) is a close approximation to constant signal-current conditions. The V_2/V_1 curve in Fig. 14.20 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. 14.30. 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. 14.20, 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

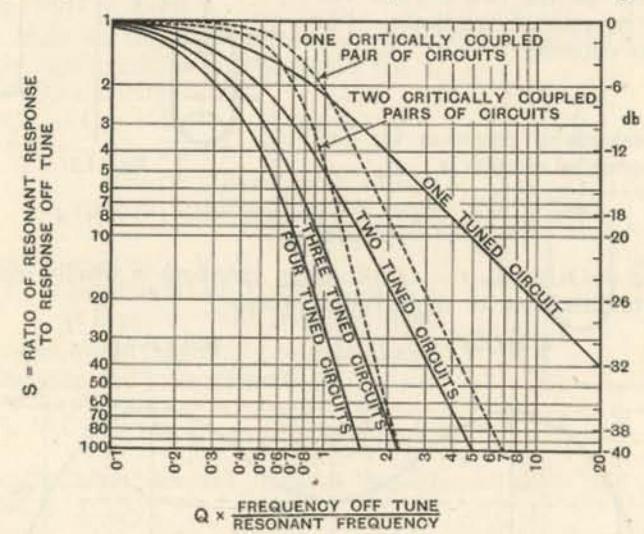


Fig. 14.30—Universal response curves, showing the extent to which the response falls off as there is departure from the resonant frequency. Comparing one critically-coupled pair of circuits with two single circuits, one can see that the steepness of the slope well away from resonance is the same, but the coupled pair gives a more level response near resonance

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 $\alpha = Qf'/f_r = 2\pi f'L/r$; N = number of tuned circuits
f_r = frequency of resonance ; 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, $A_1 \simeq Qg_m\omega L = Qg_m/\omega C = Q^2 g_m r = g_m R$.

Coupled Circuits (Fig. 14.31)

Coefficient of coupling, $k, = X_m/\sqrt{X_1 X_2}$, where X_1 and X_2 are the reactances of the two circuits and X_m is the mutual reactance. For inductive coupling

$$k = M/\sqrt{L_1 L_2}$$

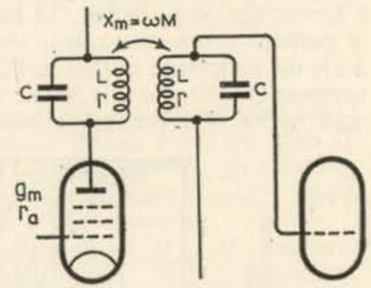


Fig. 14.31

The selectivity equation for one pair of coupled circuits is

$$S \simeq \frac{2}{1 + \beta^2}\sqrt{4\alpha^4 + 2(1 - \beta^2)\alpha^2 + (1 + \beta^2)^2/4}$$

where $\beta = Qk = X_m/r = QM/L$ when coupling is wholly inductive. Other conditions as for single tuned circuits.

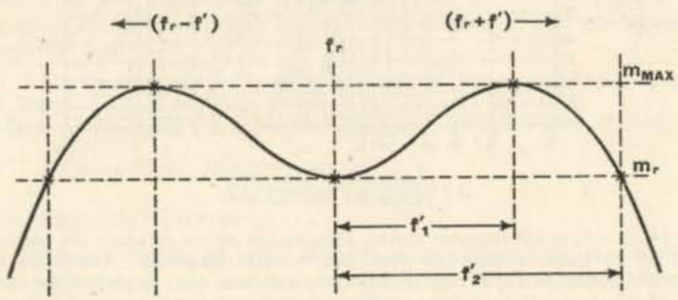


Fig. 14.32

Condition 1 $\beta = 1$ (critical or optimum coupling)
$X_m = r$

Then m_{max} at resonance $= Q/2$ and $A \simeq Qg_m\omega L/2$ $[r_a >> R]$
$S \simeq \sqrt{4\alpha^4 + 1}$, or, for N pairs of circuits, $(4\alpha^4 + 1)^{N/2}$, from which response curves can be worked out (dotted in Fig. 14.30).

Condition 2 $\beta < 1$ (under-coupling)
$m_{max} < Q/2$

Condition 3 $\beta > 1$ (over-coupling) (Fig. 14.32)
$m_{max} = Q/2$, but not at resonance

$$\text{Response at resonance, } m_r, = \frac{Q\beta}{1 + \beta^2} \quad \text{and} \quad \frac{m_{max}}{m_r} = \frac{1 + \beta^2}{2\beta}$$

$$f_1' = \pm\frac{f_r\sqrt{\beta^2 - 1}}{2Q} \qquad f_2' = \sqrt{2}f_1'$$

from which five important points can easily be calculated:

$$A \text{ (at resonance)} \simeq \frac{Qg_m\omega L\beta}{1 + \beta^2}$$

$$A \text{ (at peaks)} \simeq \frac{Qg_m\omega L}{2}$$

$$S \text{ at resonance} = (1 + \beta^2)/2\beta$$

14.20. Miller Effect

The effect of the interelectrode capacitances (Fig. 14.33) 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]

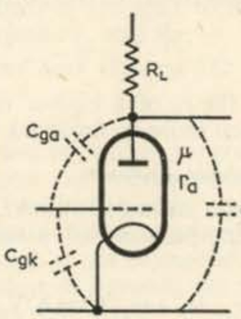


Fig. 14.33

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.

14.21. 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 in opposition to the signal, the amplification is reduced to

$$A' = \frac{A}{AB + 1}$$

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 per cent of the output is fed back in opposition ($B = 0{\cdot}06$). Then $A' = \dfrac{500}{1 + 30} = 16$.

If $AB >> 1$, $A' \simeq \dfrac{1}{B}$; that is to say, the amplification is virtually unaffected by minor changes in the amplifier itself and depends only on B, 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

$$r_a' = \frac{r_a}{\mu B + 1}$$

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.

14.22. Valve Equivalent Generator

Fig. 14.34*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

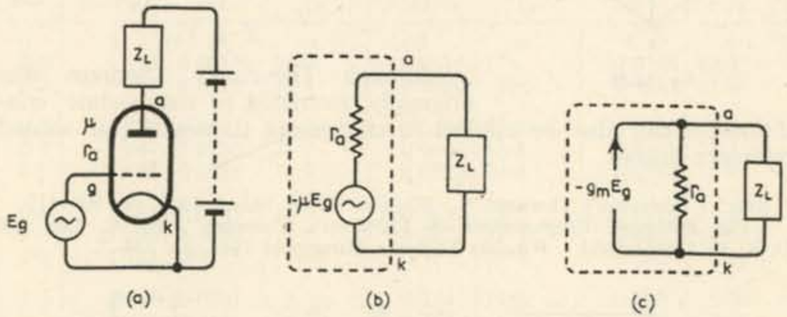


Fig. 14.34

grid voltages relative to cathode, it is necessary for the generator 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: *Wireless World*, April 1947, pp. 129–130, and April 1951, pp. 152–4.

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. 14.35 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

* "Negative Feedback and Hum." *Wireless World*, May 1946, pp. 142–5.

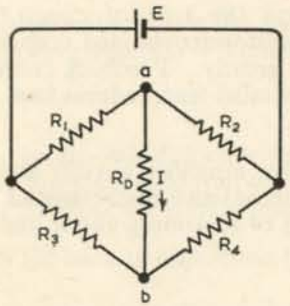


Fig. 14.35

is the " generator " resistance (R_1 and R_2 in parallel, in series with R_3 and R_4 in parallel).

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 (see second reference below).

Refs: " Thévenin's Theorem ". *Wireless World*, March 1949, pp. 109–112.
" The Extended Employment of Thévenin's Theorem ", by A. Lee and D. K. C. MacDonald. *Wireless Engineer*, November 1945, pp. 534–7.

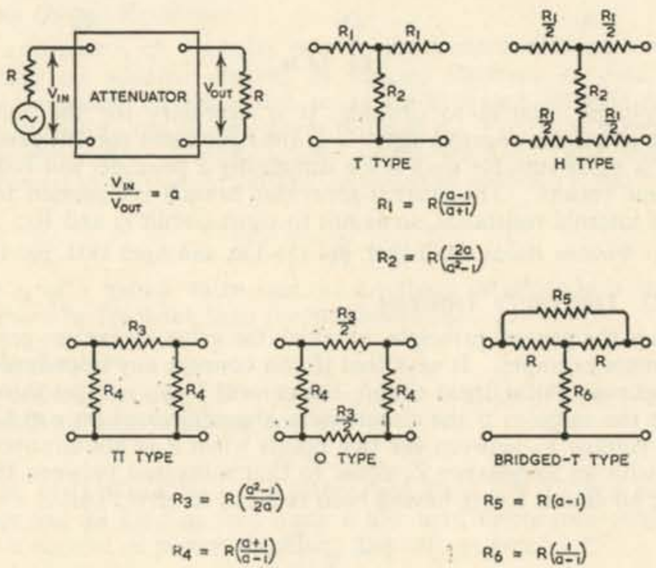


Fig. 14.36

14.24. Attenuators

Table 14.10 covers the most important types of attenuator used between equal resistances, R (Fig. 14.36). 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.

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, pp. 213–233.

14.25. Smoothing and Decoupling Filters

If the load impedance $>> 1/\omega C$, the simple RC smoother or decoupler (Fig. 14.37) is an example of Sec. 14.15. In Fig. 14.38 the same information (full-line curves) is presented in convenient form. In Fig. 14.37, V_{in} and V_{out} refer to the ripple or hum voltage. Then

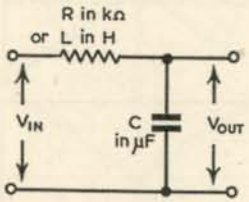


Fig. 14.37

$$\frac{V_{in}}{V_{out}} = \sqrt{\left(\frac{R}{X_C}\right)^2 + 1}.$$

If, as would normally be so, R/X_C is at least 4,

$$\frac{V_{in}}{V_{out}} \simeq \frac{R}{X_C} = 2\pi fRC \qquad [R \nless 4X_C, \text{ or } fRC \text{ in c/s – k}\Omega\text{ – }\mu\text{F} \nless \text{ say } 600]$$

In the LC filter,

$$\frac{V_{in}}{V_{out}} = \frac{X_L}{X_C} - 1 = 4\pi^2 f^2 LC - 1 \text{ (dotted curves in Fig. 14.38)}$$

Sectionalizing

If R/X_C exceeds 16 (fRC in kc/s – kΩ – μF $> 2{\cdot}5$), it is advantageous to split R and C into equal ℸ sections, cascaded. The advantage increases very rapidly with fRC:

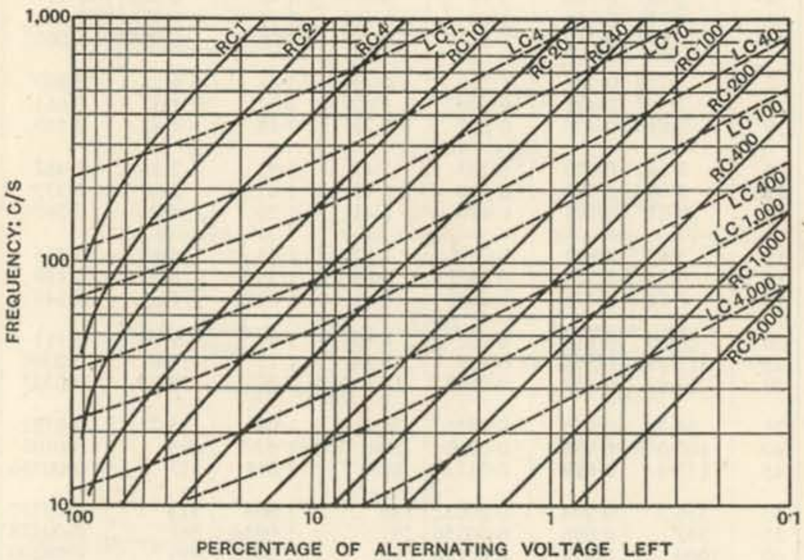


Fig. 14.38—Curves showing the effectiveness of the simple types of decoupler or hum filter (Fig. 14.37). The RC figures are C in μF multiplied by R in kΩ; LC figures are C in μF multiplied by L in H. It is assumed that the impedance of the load is much greater than that of C

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. 14.39)

With capacitance input filter, assuming current through rectifier into C_r flows as pulses at peaks of alternating voltage:

$$\text{Hum voltage} \simeq \frac{\sqrt{2}\, I}{(\omega^2 LC - 1)\omega C_r}$$

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).

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.

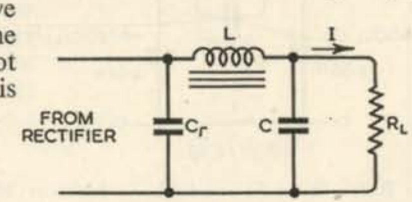


Fig. 14.39

Ref: "Smoothing Circuits". *Wireless World*, October 1949, pp. 389–393, and November 1949, pp. 418–422.

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 arms. In either case, a factor of 2 or 0·5 has to be used for Π or T section arms.

For a half section (˥ or Γ), use Table values.

For a T section, use Table values for series arms, and half Table inductance and double Table capacitance for shunt 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), and at the same time obtaining better impedance matching. For optimum matching, *m* should be 0·6 or slightly over.

E.g: Design a low-pass filter section to cut out 9-kc/s whistle and match 5 kΩ.

Obviously an *m*-derived type is required, and as no cut-off frequency is mentioned *m* will be assumed to be 0·6, so

$$0{\cdot}6 = \sqrt{1-\left(\frac{f_c}{9}\right)^2} \qquad \therefore f_c = 7{\cdot}2 \text{ kc/s}$$

$$L = \frac{mR_0}{\omega_c} = \frac{0{\cdot}6\times 5}{2\pi\times 7{\cdot}2} = 0{\cdot}066 \text{ H}$$

$$C = \frac{m}{\omega_c R_0} = \frac{0{\cdot}6}{2\pi\times 7{\cdot}2\times 5} = 0{\cdot}00266\ \mu\text{F}$$

$$C_p = \frac{1-m^2}{m\omega_c R_0} = \frac{0{\cdot}64}{0{\cdot}6\times 2\pi\times 7{\cdot}2\times 5} = 0{\cdot}00472\ \mu\text{F}$$

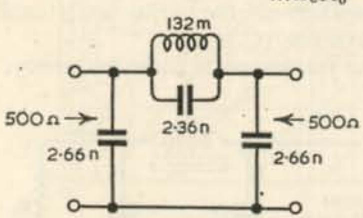
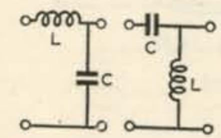
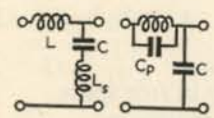
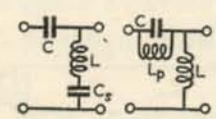
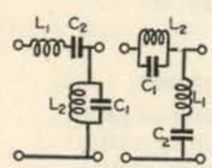


Fig. 14.40

Assuming capacitors are easier to provide than inductors, a Π section would be chosen; and following the rules for combining two half-sections gives the result shown (Fig. 14.40).

Refs: *Wave Filters*, by L. C. Jackson (Methuen).
"Filters". *Wireless World*, January 1950, pp. 25–29, and February 1950, pp. 61–65.

Table 14.13

Filter half-sections	*Component values* (R_0=input and output terminating voltage; $\omega_c = 2\pi\times$cut-off frequency; $\omega_\infty = 2\pi\times$frequency of maximum attenuation; $\omega_1 = 2\pi\times$lower cut-off frequency; $\omega_2 = 2\pi\times$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 	$L_1 = \frac{R_0}{\omega_2-\omega_1}$ $C_1 = \frac{1}{(\omega_2-\omega_1)R_0}$ $L_2 = \frac{(\omega_2-\omega_1)R_0}{\omega_1\omega_2}$ $C_2 = \frac{\omega_2-\omega_1}{\omega_1\omega_2 R_0}$

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.

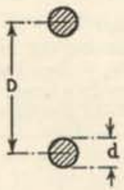


Fig. 14.41

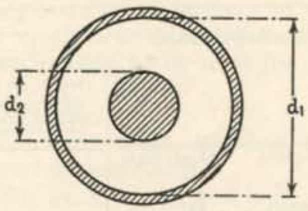


Fig. 14.42

Parallel-wire Line (section, Fig. 14.41); 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]

Coaxial Line (section, Fig. 14.42); relative permittivity of spacing, κ_s

$L \simeq 0{\cdot}46 \log_{10}(d_1/d_2)$ microhenries per metre
$C \simeq 24{\cdot}1\ \kappa_s/\log_{10}(d_1/d_2)$ picofarads per metre
$Z_0 \simeq (138/\sqrt{\kappa_s}) \log_{10}(d_1/d_2)$ ohms

Velocity of propagation (phase velocity) $= c/\sqrt{\kappa_s}$
c = velocity in free space $\simeq 3 \times 10^8$ metres/sec

Wavelength along line $= \lambda/\sqrt{\kappa_s}$ 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

$$Z_1 = Z_0 \frac{Z_2 + jZ_0 \tan 2\pi d}{Z_0 + jZ_2 \tan 2\pi d}$$

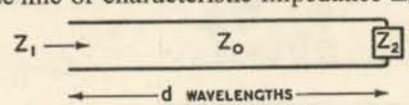


Fig. 14.43

Refs: *High Frequency Transmission Lines*, by Willis Jackson (Methuen). *Wireless World*, July 1950, pp. 251–4, and August 1950, pp. 283–6.

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.

14.29. Cathode Ray Deflection

Velocity of electron = $594\sqrt{V}$ km/sec
[V $\ngtr$ 10,000, beyond which increase in electron mass due to velocity is appreciable]

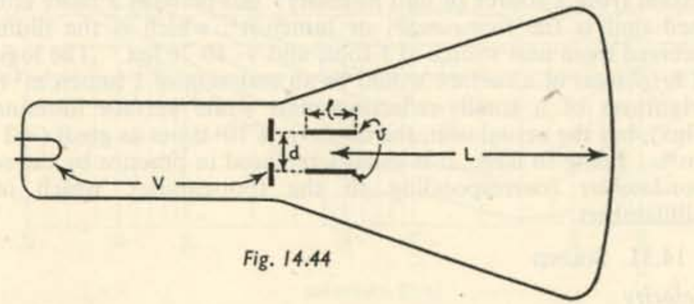


Fig. 14.44

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, gauss

Magnetic Deflection

$$D \simeq \frac{0{\cdot}3lLH}{\sqrt{V}}$$ [D, l, L, all in cm]

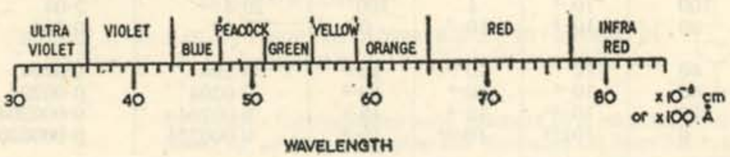
14.30. Light

Velocity

$c = 299{,}792{\cdot}5 \pm 3$ km/s (N.P.L. cavity-resonator method).

Ref: *Wireless Engineer*, G.W.O.H., April 1951, pp. 99–100.

Visible Wavelength Range (*Fig. 14.45*)



Units

Light units are based on the idea of *luminous flux*, or radiant energy, 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 intensity is an arbitrary one, the *international candle*. The total flux emitted by a source of unit intensity is 4π *lumens*, so 1 lumen is the flux flowing through unit area at unit radius from unit source. The unit of flux density, or *illumination*, equal to 1 lumen/m², is the *lux*, and is the illumination received at right angles by a surface 1 metre distant from a source of unit intensity; but perhaps a more commonly used unit is the *foot-candle*, or lumen/ft², which is the illumination received from unit source at 1 foot, and = 10·76 lux. The logical unit of *brightness* of a surface would be an emission of 1 lumen/m² (i.e., the brightness of a totally-reflecting matt white surface illuminated by 1 lux), but the actual unit, the *lambert*, is 10^4 times as great (=1 lumen/cm²). Being so large, it is usually replaced in practice by the so-called *foot-lambert* (corresponding to the foot-candle), which is 1·076 millilambert.

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/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 at 20°C and 0·76 m of Hg*	
	W/m²	*μW/cm²*	*ergs/cm²*	*dynes/cm²*	*newton/m²*
120	1	100	1000	204	20·4
100	10^{-2}	1	100	20·4	2·04
80	10^{-4}	10^{-2}	0·1	2·04	0·204
60	10^{-6}	10^{-4}	10^{-3}	0·204	0·0204
40	10^{-8}	10^{-6}	10^{-5}	0·0204	0·00204
20	10^{-10}	10^{-8}	10^{-7}	0·00204	0·000204
0	10^{-12}	10^{-10}	10^{-9}	0·000204	0·0000204

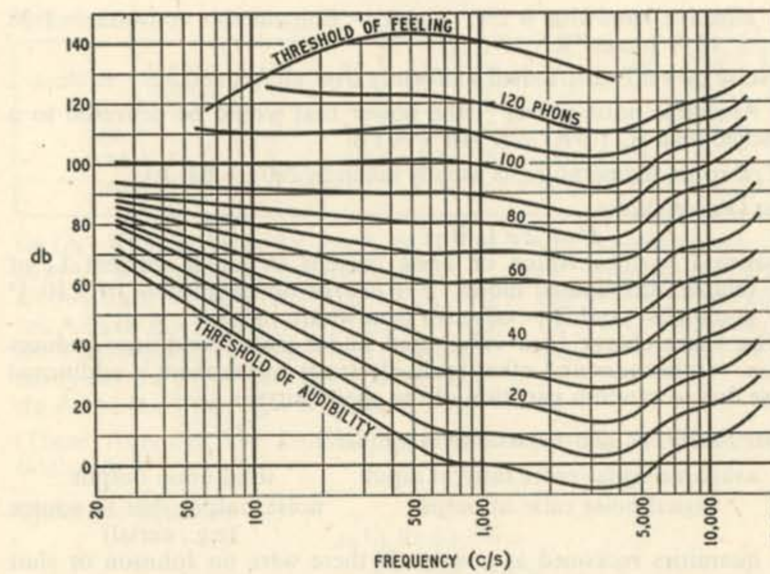


Fig. 14.46—Fletcher-Munson curves, showing the intensity of sound (in db above threshold of audibility at 1 kc/s) corresponding to loudness (in phons) as estimated by a person with normal hearing by comparison with sound at 1 kc/s

The sensitivity of microphones is usually specified as db below 1 volt per dyne/cm².

Loudness

This 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 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-Munson curves (Fig. 14.46). Sometimes, especially in America, loudness 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 is preferable.

14.32. Noise

Thermal-agitation (Johnson) Noise

$$E^2 = 4\,kTBR$$

where E = r.m.s. value of noise e.m.f. in circuit of resistance R at temperature T (Kelvin), within an

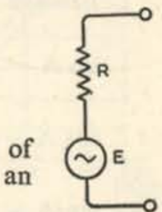


Fig. 14.47—The fluctuation noise produced by a resistance R is represented by an imaginary generator producing the same r.m.s. e.m.f. E

effective bandwith 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) $=E^2/4R = k$TB.

T, if room temperature, is usually taken as 290 (= 63° F).

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 = $1{\cdot}6\times10^{-19}$ coulombs; and B = effective bandwidth, in c/s.

The space charge in a valve used under usual conditions 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.

Ref: "Noise." *Wireless World,* May 1952, pp. 199–202, and June 1952, pp. 222–6.

14.33. Mathematical Formulae*

Circles and Spheres

r = radius; d = diameter; π = 3·1416 ...

Circumference of circle $= 2\pi r = \pi d$

Area of circle $= \pi r^2 = \dfrac{\pi d^2}{4}$

Surface area of sphere $= 4\pi r^2 = \pi d^2$

Volume of sphere $= \dfrac{4\pi r^3}{3} = \dfrac{\pi d^3}{6}$

Angles

$1 \text{ radian} = \dfrac{180^\circ}{\pi} = 57{\cdot}3^\circ$ In mathematical formulae, angles are in radians.

$\sin A = \dfrac{a}{c}$ $\quad \operatorname{cosec} A = \dfrac{c}{a}$

$\cos A = \dfrac{b}{c}$ $\quad \sec A = \dfrac{c}{b}$

$\tan A = \dfrac{a}{b}$ $\quad \cot A = \dfrac{b}{a}$

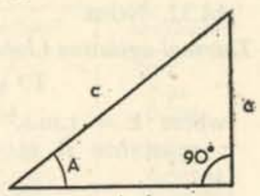


Fig. 14.48

* For symbols, see p. 363.

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

$\sin(A\pm B) = \sin A\cos B \pm \cos A \sin B$
$\cos(A\pm B) = \cos A\cos B \mp \sin A\sin B$
$\sin A + \sin B = 2\sin\frac{1}{2}(A+B)\cos\frac{1}{2}(A-B)$
$\cos A + \cos B = 2\cos\frac{1}{2}(A+B)\cos\frac{1}{2}(A-B)$
$\sin A\sin B = \frac{1}{2}[\cos(A-B) - \cos(A+B)]$
$\cos A\cos B = \frac{1}{2}[\cos(A+B) + \cos(A-B)]$
$\sin A\cos B = \frac{1}{2}[\sin(A+B) + \sin(A-B)]$

$\sin^2 A + \cos^2 A = 1$
$\sin^2 A = \frac{1}{2}(1-\cos 2A)$
$\cos^2 A = \frac{1}{2}(1+\cos 2A)$
$\sin 2A = 2\sin A\cos A$
$\cos 2A = \cos^2 A - \sin^2 A$

(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

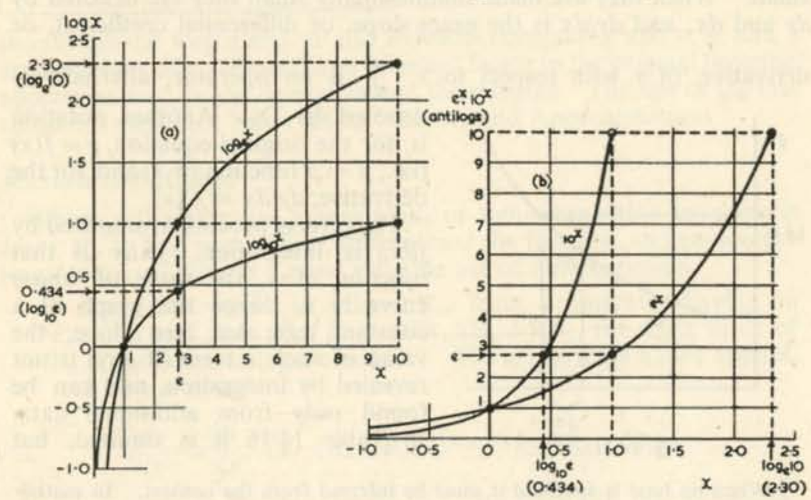


Fig. 14.49—(a) Logarithmic and (b) exponential curves, for 10 and e. b is the same as a turned over a diagonal line

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. 14.49*a*.)

The following *exponential functions* (Fig. 14.49*b*) are much used in electrical theory:

$$e^x = 1 + x + \frac{x^2}{2!} + \frac{x^3}{3!} + \frac{x^4}{4!} + \cdots$$

$$e^{jx} = \cos x + j \sin x$$

(represents a vector of unit length, at angle x radians)

$$e^{-jx} = \cos x - j \sin x$$

$$\sin x = \frac{e^{jx} - e^{-jx}}{2j} \qquad \cos x = \frac{e^{jx} + e^{-jx}}{2}$$

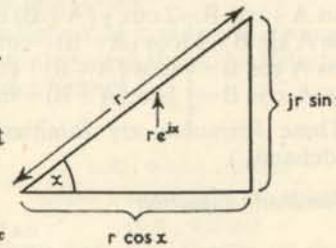


Fig. 14.50

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 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".

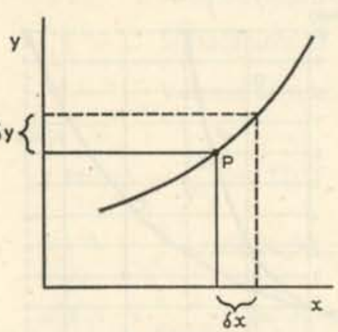


Fig. 14.51

* 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".

Table 14.16

y	$\frac{dy}{dx}$	$\int y.dx$	$\frac{x}{y}\frac{dy}{dx}$
a	0	ax	0
x^n	nx^{n-1}	$\frac{x^{n+1}}{n+1}$ ($n \neq -1$)	n
$ax^n + b$	nax^{n-1}	$\frac{ax^{n+1}}{n+1} + bx$ ($n \neq -1$)	$\frac{n}{1+b/ax^n}$
e.g: ax $n=1$	a	$ax^2/2$	1
ax^2 $n=2$	$2\,ax$	$ax^3/3$	2
ax^3 $n=3$	$3\,ax^2$	$ax^4/4$	3
$a\sqrt{x}$ $n=\frac{1}{2}$	$a/2\sqrt{x}$	$\frac{2}{3}\,ax\sqrt{x}$	$\frac{1}{2}$
a/x $n=-1$	$-a/x^2$	$a \log_e x$	-1
a/x^2 $n=-2$	$-2a/x^3$	$-a/x$	-2
$a/\sqrt{x}$ $n=-\frac{1}{2}$	$-a/2x\sqrt{x}$	$2a\sqrt{x}$	$-\frac{1}{2}$
$\sin(ax+b)$	$a\cos(ax+b)$	$-\frac{1}{a}\cos(ax+b)$	$ax/\tan(ax+b)$
$\cos(ax+b)$	$-a\sin(ax+b)$	$\frac{1}{a}\sin(ax+b)$	$-ax\tan(ax+b)$
$\log_e(ax+b)$	$\frac{a}{ax+b}$	$\left(x+\frac{b}{a}\right)\log_e(ax+b) - x$	$\frac{ax}{(ax+b)\log_e(ax+b)}$
e^{ax}	ae^{ax}	e^{ax}/a	ax
$u+v$	$\frac{du}{dx} + \frac{dv}{dx}$	$\int u.dx + \int v.dx$	
uv	$u\frac{dv}{dx} + v\frac{du}{dx}$		
$\frac{u}{v}$	$\left(v\frac{du}{dx} - u\frac{dv}{dx}\right)/v^2$		
$f(u)$	$\frac{dy}{du} \cdot \frac{du}{dx}$		

E.g.: Differentiate $\frac{x^2}{x+1}$

This is covered by the procedure (on the left) for $\frac{u}{v}$: $\frac{(x+1) \times 2x - (x^2 \times 1)}{(x+1)^2} = \frac{x(x+2)}{(x+1)^2}$

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

$$v\frac{du}{dx} = u\frac{dv}{dx}; \text{ i.e. } (R_L+R)^2\,E^2 = E^2R_L(2R_L+2R),$$

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 $\simeq \frac{\delta x}{x}\cdot\frac{x}{y}\cdot\frac{dy}{dx}$. The last column in Table 14.16 gives $\frac{x}{y}\cdot\frac{dy}{dx}$ for various functions, and represents the approximate ratio of $\delta y/y$ to $\delta x/x$, or the percentage change in y for 1 per cent 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 per cent increase in C, f_0 *de*creases approximately $\frac{1}{2}$ per cent. 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.

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. An amplifier with a voltage gain of 20 would (up to its overload point) give a signal output of 20 V for every 1 V at the input. With another stage having a gain of 15, the total gain

would be 15 *times* 20, = 300. The same principle applies to power gain.

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. 14.52 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 $2\frac{1}{2}$ times 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 represents gives more uniform amplification. But in actual fact all three curves indicate identical frequency characteristics.

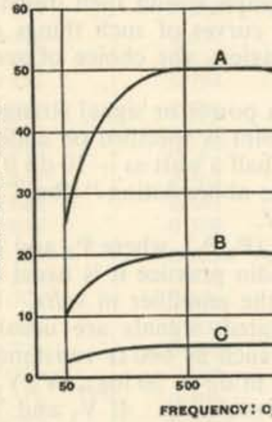


Fig. 14.52—Three characteristic curves plotted on an ordinary magnification or gain scale. The general appearance of the curves leads to quite wrong conclusions

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

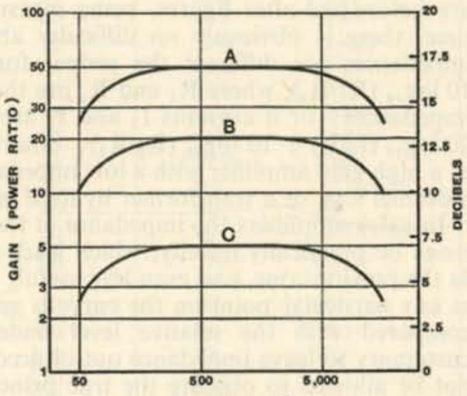


Fig. 14.53—When the curves of Fig. 14.52 are replotted on a logarithmic gain scale, which is the same as a linear decibel scale, the true proportions appear

things that actually are equal look equal. This object is attained if gain is plotted on logarithmically-divided paper. Fig. 14.53 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. 14.53 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 " 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 output power of a receiver is observed to rise from 46 to 58 mW as 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}	$3{\cdot}16\times10^{5}$
10^{-6}	10^{-12}	120	10^{12}	10^{6}

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:

Number of nepers $= \log_e (I_2/I_1)$ or $\frac{1}{2} \log_e (P_2/P_1)$

Given equal impedance:

1 n = 8·686 db   1 dn (decineper) = 0·8686 db
1 db = 0·1151 n   1 db = 1·151 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 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

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

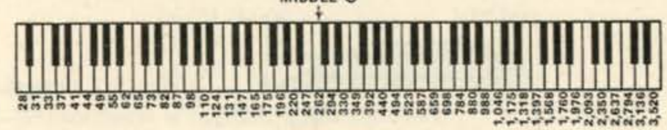


Fig. 14.54—Piano scale showing the frequencies to which the keys are tuned. The usually published scales based on C = 256 are misleading. Frequencies of black keys can be obtained by multiplying the frequency of the next white key below by 1·059

of the notes are shown in Fig. 14.54. 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.

14.36. Frequency Allocations

International Allocations at Atlantic City, 1947

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 and Africa). Region 2 is N. and S. America, and Region 3 Asia 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	R 7150–7300 kc/s	S 3500–3800 kc/s	2498–2502 kc/s
R 160–255	9500–9775	7000–7100	4995–5005
SR 255–285	11700–11975	S 7100–7150	9995–10005
R 525–535	15100–15450	14000–14350	14990–15010
535–1605	17700–17900	21000–21450	19990–20010
S 2300–2495	21450–21750	28000–29700	24990–25010
SR 2495–2498	25600–26100	144–146 Mc/s	
S 3200–3400	R 41–68 Mc/s (I)	S 420–450	
SR 3950–4000	87·5–88 } (II)	SR 450–460	
SR 4750–4850	88–100* } (II)	1215–1300	
S 4850–4995	R 174–216 (III)	2300–2450	
S 5005–5060	470–585 (IV)	5650–5850	
5950–6200	610–940 } (V)	10000–10500	
SR 7100–7150	R 940–960 } (V)	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 Mc/s not available for broadcasting in U.K.			

British Television Channels

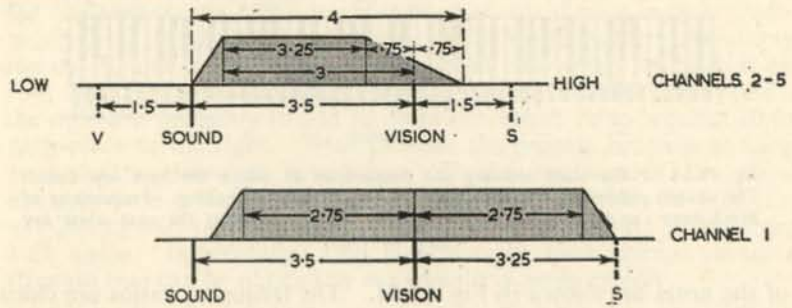


Fig. 14.55

Table 14.20

Channel	*Sound*	*Vision*
1 (London)	41·5 Mc/s	45·0 Mc/s
2 (Holme Moss)	48·25	51·75
3 (Kirk o'Shotts)	53·25	56·75
4 (Sutton Coldfield)	58·25	61·75
5 (Wenvoe)	63·25	66·75

Fig. 14.55 shows the channel spacings in Mc/s, and the ideal frequency responses for vision. The dotted lines indicate adjacent-channel frequencies.

International television standards are tabulated in *Wireless World*, August 1952, p. 296.

14.37. STANDARD FREQUENCIES

B.B.C. Transmitters

The following transmitters of the British Broadcasting Corporation are guaranteed to maintain their nominal frequencies within 1 in 10^6:

Droitwich	200 kc/s
GRO	6,180
GSB	9,510
GSV	17,810

Droitwich and 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 $\pm$ 5 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 $\pm$ 5 in 10^8.

G.P.O. Transmitter

The frequency of the Rugby transmitter GBR is 16 kc/s $\pm$1 in 10^6.

U.K. Standard-frequency Transmissions

These signals are radiated daily from G.P.O. transmitters at Rugby, call sign MSF, with a power of 10 kW, to the following schedule:

05·44–06·15 G.M.T.	5 Mc/s
06·29–07·00	10 Mc/s
10·29–11·30 } 14·29–15·30 }	60 kc/s

Each transmission is divided up as follows:

Minutes 0–1: call sign in slow morse; speech announcement
1–6: carrier modulated at 1 kc/s
6–11: 1-sec timing pulses
11–15: carrier unmodulated
15–16: call sign and announcement; then 15-minute cycle (1–16) repeated.

The frequencies are maintained within $\pm$ 2 parts in 10^8 of the nominal values. Inquiries or comments should be addressed to the Director, National Physical Laboratory, Teddington, Middlesex.

American Bureau of Standards Standard-frequency Transmissions

A continuous 24-hour service from the Bureau's radio station WWV near Washington, D.C., comprises standard radio and audio frequencies, as listed in Table 14.21. At least one, and often several,

Table 14.21

Carrier frequency, $\pm$2 in 10^8 (Mc/s)	*Power* (kW)	*Modulation frequency,* $\pm$2 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
30 35	0·1 0·1	440

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 $\pm$ 2 in 10^8. WWV in morse.
29–30, 59–60	Station announcement by voice
19–20, 49–50	Radio propagation signals, as on p. 416

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.

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

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 only one of the patterns can be made stationary on the screen at a time.

14.38. 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.

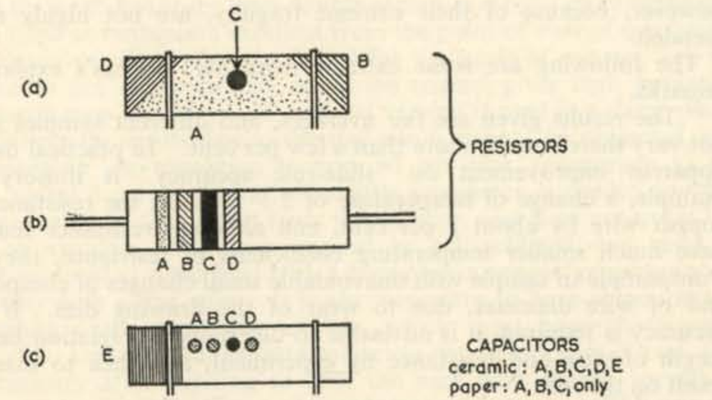


Fig. 14.56—Standard colour markings of resistors and capacitors; for key, see Table 14.23. An alternative to the spot C in a is a band. A salmon-pink fifth band in b indicates Grade I

Table 14.23

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^{-1}	—	± 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 (B.S.I.).

14.39. 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 per cent, 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 per cent 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. 14.57, not as in Fig. 14.58. Although Fig. 14.58 is possibly nearer the truth when a wire winding machine is used, experience shows that when wire is wound by hand in a lathe, 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.

Fig. 14.57

Fig. 14.58

"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 per cent, manganese 12 per cent, nickel 4 per cent), 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 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."

Table 14.24

S.W.G.	Copper: Ohms per km	Copper: *Metres per 1 oz reel D.S.C.	Copper: *Metres per 1 lb reel D.C.C.	Constantan (or Eureka): Cm per ohm	Constantan (or Eureka): Ohms per 1 oz reel D.S.C.	Constantan (or Eureka): Ohms per 1 lb reel D.C.C.	Manganin: Cm per ohm	Manganin: Ohms per 1 oz reel D.S.C.	Manganin: 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: Ohms per km	Copper: *Metres per 1 oz reel D.S.C.	Copper: *Metres per 1 lb reel D.C.C.	Constantan (or Eureka): Cm per ohm	Constantan (or Eureka): Ohms per 1 oz reel D.S.C.	Constantan (or Eureka): Ohms per 1 lb reel D.C.C.	Manganin: Cm per ohm	Manganin: Ohms per 1 oz reel D.S.C.	Manganin: Ohms per 1 lb reel D.C.C.	S.W.G.
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 14.25

S.W.G.	Turns per cm length of single-layer coil: D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel	Turns per sq. cm. section of solid coil: D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel	Diameter (in) (bare)	Cross-section (sq. mm)	Diameter (mm) (bare)	S.W.G.
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: D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel	Turns per sq. cm. section of solid coil: D.S.C.	D.C.C.	E. & S.S.	E. & S.C.	Enamel	Diameter (in) (bare)	Cross-section (sq. mm)	Diameter (mm) (bare)	S.W.G.
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

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The following are printed in italics:
(1) *Names of persons whose work is referred to*
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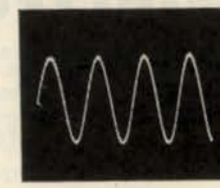

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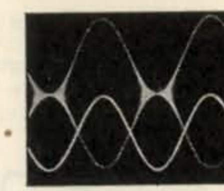

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