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AN OUTLINE OF THE PRINCIPLES UPON WHICH RADIO RECEIVER TEST AND MEASURING INSTRUMENTS OPERATE

BY

W. H. CAZALY

LONDON
SIR ISAAC PITMAN & SONS, LTD.
1945

SIR ISAAC PITMAN & SONS, LTD.
PITMAN HOUSE, PARKER STREET, KINGSWAY, LONDON, W.C.2
THE PITMAN PRESS, BATH
PITMAN HOUSE, LITTLE COLLINS STREET, MELBOURNE
UNITEERS BUILDING, RIVER VALLEY ROAD, SINGAPORE
27 BECKETTS BUILDINGS, PRESIDENT STREET, JOHANNESBURG

ASSOCIATED COMPANIES
PITMAN PUBLISHING CORPORATION
2 WEST 45TH STREET, NEW YORK
205 WEST MONROE STREET, CHICAGO
SIR ISAAC PITMAN & SONS (CANADA), LTD.
(INCORPORATING THE COMMERCIAL TEXT BOOK COMPANY)
PITMAN HOUSE, 381-383 CHURCH STREET, TORONTO

BOOK PRODUCTION WAR ECONOMY STANDARD

THIS BOOK IS PRODUCED IN COMPLETE CONFORMITY WITH THE AUTHORIZED ECONOMY STANDARDS

MADE IN GREAT BRITAIN AT THE PITMAN PRESS, BATH
D5—(T.286)

ADVICE TO AMATEUR CONSTRUCTORS

As a keen constructor himself, the author has every sympathy with readers who want to build their own instruments. He warns them, however, that a very clear understanding of fundamental a.c. and radio theory and principles is necessary for the design of even apparently simple apparatus. He regrets that he cannot himself supply them with explicit constructional data, as regards either mechanical structure or component values. This is because

(1) The author is not at liberty to give specific circuit details of commercial instruments.

(2) Design data for *all* the instruments mentioned are to be found in a large number of technical books and periodicals if people take the trouble to search for themselves—which is far better for their own radio education than being told exactly where to look.

(3) Construction involves not only buying and assembling the right components, but adjustment and calibration of the instrument as a whole—and this is usually much more difficult than mere assembly and requires skill and knowledge that cannot be imparted either in a book of this nature or in correspondence.

It is the hope of the author, however, that readers *will* attempt construction themselves, because this, in parallel with voracious reading of technical literature, is by far the best method of improving one's knowledge of radio theory and technique. The school of practical experience still charges extremely high fees, but in radio they are worth paying. Below is a short list of literature that the reader should peruse.

FUNDAMENTAL THEORY

Admiralty Handbook, Vols. 1 and 2.
Terman's *Radio Engineering* (McGraw-Hill).

INSTRUMENT THEORY AND TECHNIQUE

Scroggie's *Radio Laboratory Handbook* (Iliffe).
Terman's *Measurements in Radio Engineering* (McGraw-Hill).

Hartshorn's *Radio Frequency Measurements and Bridge and Resonance Methods* (Chapman and Hall).

Radio Designer's Handbook (Amalgamated Wireless (Australasia), Ltd., distributed in Great Britain from *Wireless World* Office, Iliffe & Co.).

A number of articles on the principles and construction of various instruments have appeared in past numbers of *Wireless World* and of *Electronics*, both now monthlies. It is largely a matter of luck, however, whether one can obtain any back copies and one must specify the back number needed.

There are, of course, many other sources of information.

ACKNOWLEDGMENT

THE illustrations, and all but the first chapter of this book, appeared originally with very slight alterations as articles in the *Wireless World*, the Editor of which journal, Mr. H. F. Smith, the author wishes sincerely to thank for his help and kindly encouragement in the preparation of the original articles for publication.

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RADIO SERVICE TEST GEAR

CHAPTER I

PERFORMANCE

THE time has long passed since the performance of a radio receiver for the reproduction of sound was judged by the professional by the subjective personal impressions it made on a listener. Radio receivers, in fact, are no longer "powerful" or "mellow in tone" to properly trained servicemen and other radio technicians. By "powerful" is probably meant either high absolute sensitivity in microvolts input for a small given output in watts, or a large wattage output for a given, not necessarily small, input. By "mellow tone" is probably meant either unusually quiet background noise, free from interference, valve hiss, residual hum, etc., or, as is far more common, a cut-off in the region of 2000 c/s, which removes very desirable brilliance as well as shrill noises. Such vague, non-technical words as "loud," "powerful," "clear," etc., are of little interest to a radio technician or serviceman of adequate competence and professional status. It is, of course, permissible for the salesman and advertiser to talk in the language of the layman about performance—but to the trained technician, such language means very little.

The performance of anything is what it *does*—not what emotions or affections it rouses. What it does can, in the vast majority of cases, be measured to a high degree of accuracy by impersonal methods and instruments. Radio receivers are no exception. Since the performance of a receiver is a complex thing, it can be analysed into various components, each of which can be tested and measured separately.

Though there is not as yet finality in the acceptance of universal standards and forms of performance of radio sound receivers, it is by now possible to obtain a good idea of the merit of a receiver by subjecting it to certain very common, almost standard, tests. The results of these tests can be expressed in terms of cold figures that tell the technician much more than volumes of effective description. To carry out these tests—and most of them should be matters of routine in professional service work—certain forms of instruments are required to produce and measure the effects of definite, controllable, and precisely known conditions.

The performance of a radio sound receiver may be divided into, roughly, the following parts or aspects—

First, there is *Sensitivity*, or Overall Sensitivity. This is the ability of the receiver to use the alternating potentials developed across its input (aerial and earth) terminals for the production of output in the form of wattage available at the output terminals. It is usual to standardize the output (50 mW is a very common standard output, in a 400Ω resistive load at 400 c/s) and to measure the input potential injected at the aerial and earth terminals in terms of the carrier frequency, its amplitude, and the frequency and the depth (percentage) of modulation. An output meter to measure the output, and a standard signal generator to produce the required input, are the two primarily necessary instruments to make measurements of the sensitivity of a receiver.

Secondly, there is *Selectivity*. This is the ability of a receiver to respond far more to an input at a certain given frequency than to inputs at any other frequency. Plainly, an instrument is required (a signal generator) that can produce inputs at a number of test frequencies and an output meter that can record the resultant outputs.

Both sensitivity and selectivity can be further analysed. The sensitivity of a receiver depends on the *gain* contributed by each stage of amplification. This *Stage Gain* is another factor that can be measured by means of the signal generator and the output meter. The selectivity of a receiver depends, in a superhet, partly on the characteristics of the R.F. tuned circuits and partly on those of the band-pass coupled I.F. tuned circuits, each of which can be separately examined. The ability of the R.F. plus I.F circuits to select only the wanted input from others near to it in frequency is called *Adjacent Channel Selectivity*; the ability to reject (i.e. fail to amplify appreciably) inputs at frequencies equal to the wanted frequency plus twice the I.F. is called *Second Channel*, or *Image, Selectivity*; and the ability to reject inputs at the actual I.F. is called *I.F.*, or *Breakthrough, Selectivity*. Further, although the receiver must reject adjacent channel inputs, it must respond well to a narrow band of frequencies on either side of

the wanted carrier input, in the interests of faithful reproduction of the original intelligence transmitted; this, called the *Band-pass Response*, is in the nature of a special kind of selectivity. All these different forms of selectivity are capable of precise measurement, and must be measured if judgment of performance is to be more than a vague subjective impression without technical value.

Thirdly, there is *Overall Acoustic Response*. This is the ability of the receiver as a whole to transform the modulation of the carrier wave into audio frequency output free of the various forms of distortion. The characteristics of all the stages play their parts in this aspect of performance. If the selectivity of the R.F. circuits is excessively sharp or peaky, the higher audio frequencies, which provide the outside edges of the side-bands, are lost, giving unpleasantly muffled output in terms of sound. The same may happen in the I.F. circuits if, instead of being flat-topped, their band-pass response curves are sharply peaked. If, instead of being sharply peaked in the middle of that response curve, the characteristic has a pronounced dip, with peaks on either side, or lop-sided peaks ("rabbit's ears" in an oscilloscope), the higher audio frequencies will be excessively and unevenly amplified, giving rise to distressing distortion. In the A.F. stages, undesired resonant conditions at certain frequencies may be present, and incorrect operating voltages. In tracing the sources of distortion, a beat frequency oscillator as well as a signal generator and output meter are almost necessities.

Fourthly, there is the *A.F. Characteristic*. In regard to this aspect of performance, into which come mainly the stages that amplify after detection, the receiver has to be considered primarily as an A.F. amplifier, to test which the B.F.O. and output meter are essential.

Fifthly, there is that important accessory circuit called *A.V.C.* This is really in the nature of an automatic brake that is released if the input intensity decreases and allows reserve amplification to come into play, and also throttles down amplification if the input increases above a certain level. To test its operation, the signal generator is needed to inject known inputs of widely varying amplitude, and the output meter to measure the resultant variations in output. A poor A.V.C. characteristic soon becomes noticeable, although few realize its benefits while it is good.

It would be possible to add considerably to this list of the aspects or parts of receiver performance that can be accurately measured. Many of them, however, are mainly of interest to the designer or laboratory research worker rather than to the technical serviceman. But the foregoing are the barest minimum of parts of performance that should be checked before anything technically adequate can be stated about the merit of a receiver.

It is beyond the scope of this book even to outline the actual methods of making such tests and measurements, let alone all the others that are possible in a properly equipped workshop. Not only the fundamental principles must be understood, but, with the enormous variety of makes and types of sound receiver now on the market, there is a very wide variation in specific details of methods. The fundamental principles have to be learnt in the course of technical training, and then they have to be applied intelligently to individual cases. Instruments of good accuracy and high constancy—and constancy is perhaps the more important quality—are the tools by which principles can be applied. Without the knowledge of principles, the best of instruments can be almost valueless. With that knowledge, a comparatively modest outfit of good instruments can be made to reveal, by ingenious and scientific methods of use, almost everything necessary to judge practical performance and, if performance is not up to normal, to trace the reason to its source.

A knowledge of his tools, their possibilities and their limitations, is one of the prime assets of a technician. The following chapters will introduce the reader to the basic principles upon which operate a few of the almost standard tools, or instruments, of the competent serviceman or serious amateur.

CHAPTER II

STANDARD SIGNAL GENERATORS AND TEST OSCILLATORS

THE differences between a service-bench test oscillator costing a few pounds and a standard signal generator costing £100 are ones of refinement rather than of principle. A standard signal generator—known as an S.S.G. for short—is a device that produces, across a known impedance, alternating voltages that resemble closely the radio or intermediate frequency voltages that are handled by a receiver; these "carriers" at R.F. or I.F. may be modulated or not as desired, and their frequencies, magnitudes, and waveforms, and the frequency and depth of modulation, are accurately known and precisely controllable. These artificially produced "signals" can be used to measure the performance of a receiver in definite units and terms that are independent of the personal feelings and abilities of the operator.

A device of such capabilities cannot be altogether simple. In a highly developed form, indeed, the S.S.G. may be quite complex, for it may then consist of several instruments, each of high precision and some complexity, specially designed for working in combination; the price of such a piece of apparatus might run into thousands of pounds, and for that an accuracy of the order of a few parts in several millions might be obtained. Yet the basic principles would be comparatively simple, and not really different from those employed by the humbler instruments found in thousands of service workshops. It is not proposed here to deal with high-precision apparatus of that nature. Nor is it proposed to deal with the lowest grades of cheap or home-made test oscillators. For general purposes in design laboratories and on the work benches of serious experimenters, an S.S.G. of an intermediate order of precision and complexity is very often quite satisfactory; while even the test oscillator may, in the hands of an intelligent man and if it is soundly constructed, provide *comparative* data of value quite equal for many extremely practical purposes (such as servicing) to that offered by its more expensive cousin the S.S.G.

Dealing with the S.S.G. first, Fig. 1 is a block diagram of the essential parts of a system for measuring the main aspects of a receiver's performance. Each of the four main parts may be considered for the purpose of study as an autonomous device or instrument. The four of them—the modulator, the R.F. oscillator, the R.F. output monitor, and the attenuator with its dummy aerial terminating impedance—are designed to work together in combination; each is separately and thoroughly screened, and as a whole they are enclosed in a large and substantial screening case. This screening is most

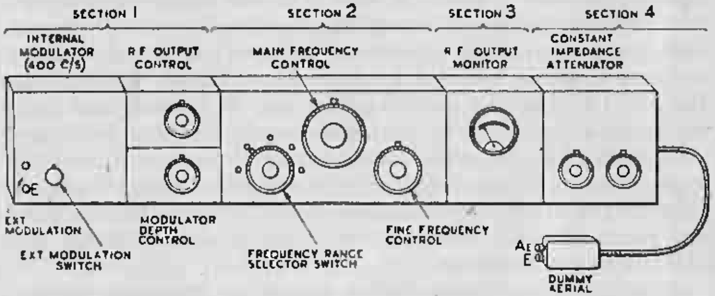


Fig. 1

The essential parts of a standard signal generator. Each part constitutes practically an instrument on its own, but all are designed to work in co-operation with each other. Modulation from an external source may be injected through the terminals marked "Ext. Modulation"

important. Without it, the system is liable to radiate stray fields of unknown and uncontrollable intensity that hopelessly invalidate the data obtained with sensitive receivers that respond to these stray fields as well as to the terminal voltages at the S.S.G. output. It may be emphasized that the foregoing states the minimum requirements for an instrument worthy of the name of *standard* signal generator; if an instrument does not possess all these features, or possesses them only in crude form incapable of precise and quantitatively known control, it may be merely an elaborate, expensive, and handsome test oscillator.

Frequency Control

The R.F. oscillator, in an S.S.G., is often the least complex and most easily designed and constructed part; no very high order of inherent long-period constancy is demanded of either

the frequency or the magnitude of the output, because both can, if necessary, be accurately measured or monitored by, respectively, an external precision wave-meter and the internal R.F. output monitoring device. It may consist merely of a robustly made valve oscillator with good, low-loss components selected for the constancy of their electrical properties. The "tuning dial," or frequency control, is usually calibrated in kilocycles and megacycles per second, and for most purposes the maker's calibration is sufficiently accurate as it stands. But an all-important adjunct to this main frequency control is a really well-made fine control that will enable the frequency of a signal to be changed or adjusted by small amounts, of the order of a few hundred or thousand c/s in several megacycles. This fine control usually takes the form of some mechanical adjustment, either as a vernier on the main control or by the movement of the stator vanes of the tuning condenser.

The R.F. output monitoring device may consist either of a valve voltmeter that records the voltage developed by the oscillator across the input impedance of the attenuator, or of a thermo-couple (for instance, a vacuo-junction) meter that registers the current through, and, therefore, indirectly the voltage across, the attenuator.

Fig. 2 shows the essential parts of the R.F. oscillator section and the R.F. output monitor, together with a practical method of controlling the R.F. output of the oscillator, in this case by varying the anode voltage by a series resistance, across which I.R. drop occurs with the passage of I_a. The procedure usually adopted is to adjust the R.F. output by varying the value of this anode series resistance until the R.F. output monitor pointer comes to a "set-up" mark on its dial; it is then assumed that a definite predetermined voltage is being developed across the attenuator input. The voltage at the anode of the R.F. oscillator will be the full voltage of the power pack less the voltage drop across the series resistance, which depends on the I_a of the R.F. oscillator. Since this I_a varies with the intensity of oscillation, and this, in turn, varies with the Q of the oscillator, which will not be the same for all frequencies and L/C ratios, the R.F. output may vary with the setting of the S.S.G. frequency control, but it can always be brought to the same value *at the input of the attenuator* by adjustment of the series anode resistance—or the V_a of the R.F. oscillator.

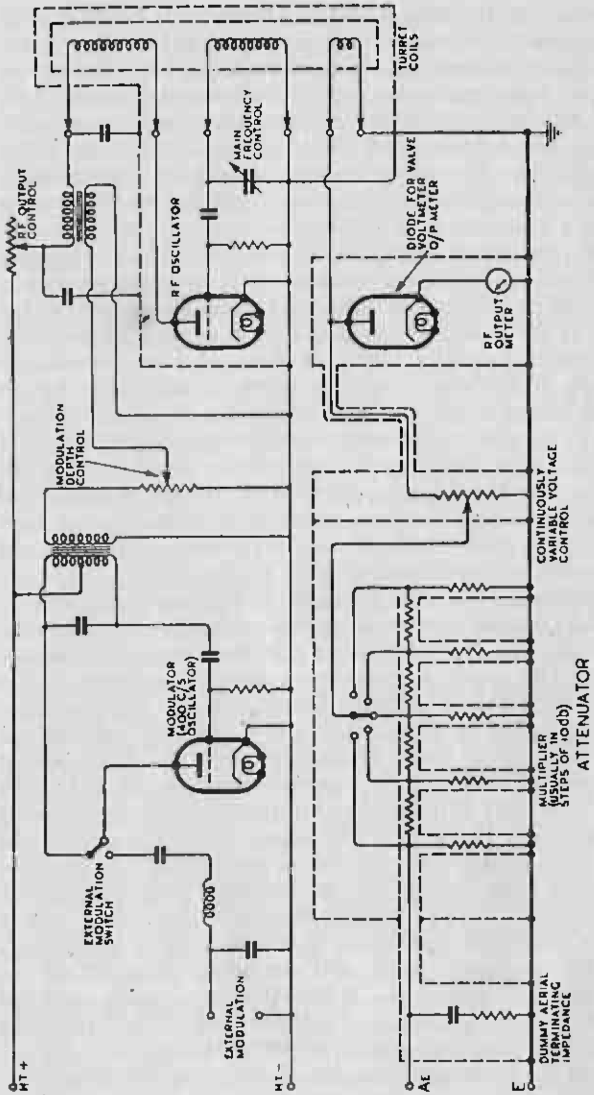


Fig. 2

Skeleton circuit of a typical S.S.G. of general-purpose type. Note the thorough screening of each section separately, indicated by dotted line

Fig. 2 also shows in skeleton form a type of attenuator much used in S.S.G.'s for general purposes. It is not of truly constant impedance, for which a more elaborate and expensive form would be required, but the ratio between the variation of voltage across its output terminals and the variation of its input impedance can be made, by suitable design, very high. For example, a change of terminal output voltage of 10,000 times may be accompanied by a change of attenuator input impedance (towards the oscillator) of only two times, and as low-impedance coupling between the attenuator and the oscillator is usually adopted, the effect on the oscillator frequency may be negligible or easily allowed for. The design of constant impedance R.F. attenuators is a very lengthy subject, but it may here be mentioned that the elements of the network are usually resistive and are mounted in a casting of brass having compartments for the elements that effectively screen them from each other but are of sufficient dimensions to keep their capacity to earth to a reasonably small amount; in this way the attenuator is made practically purely resistive and almost independent of frequency variations of the signal it handles up to some 15 or 20 Mc/s. It should be obvious that the whole accuracy and value of the S.S.G. depend upon the precision of control of the magnitude of the test signal sent out; hence the attenuator and R.F. output monitor are the most important parts of the whole apparatus. Unless they are really well designed and constructed, it is impossible to do more than obtain a rough idea of the magnitude of the output terminal voltage, which is not good enough for quantitative measurements of performance.

Modulation

Since the performance of receivers is largely stated in terms of the audio output, it is essential that the R.F. carrier of the test signal should be modulated. This may be done, in an S.S.G. of the order of cost and complexity under discussion, in two ways—either by an internal modulator consisting of a fairly simple regenerative A.F. transformer-coupled oscillator producing anode modulation of the R.F. oscillator at a common standard frequency of 400 c/s, or by the injection of a modulating voltage from an external source such as a beat frequency oscillator or similar A.F. source. Further, it is essential for many tests that the depth of modulation be fairly accurately known. For the latter purpose, some control of the modulating voltage injected is necessary, and means must be provided for measuring or obtaining definite modulation depth data. In Fig. 2 is shown one way of securing this facility.

It may be best, at this point, to study the behaviour of the various parts of an S.S.G. by considering what can be done with and by it in some actual practical case of use. It will be assumed, therefore, that it is desired to ascertain (1) the overall absolute sensitivity of a receiver, and (2) the adjacent channel response at 5 kc/s off-tune at, say, 1 Mc/s. The procedure will be roughly—varying, of course, with the make and type of S.S.G. being used—as follows.

Measurement Procedure

First, the receiver is set going with an output meter connected across the correct load of the output stage, to indicate when the generally accepted standard output of 50 mW is obtained. Next, the S.S.G. is set up. For this, the internal modulation is switched on, the modulation depth control set to zero, and the R.F. output control set to minimum. While the S.S.G. is warming up, the frequency control is set to the 1-Mc/s mark, with the fine control at zero, and the attenuator is set to the calibration point—say 10 μvolts—on its control at which it is expected that the receiver will produce approximately a 50-mW A.F. output. The dummy aerial at the end of its standard lead is connected to the aerial and earth terminals of the receiver through a condenser of the order of 0·001 μF. Now the R.F. output control is turned up until the output monitoring meter pointer reaches the "Set R.F." mark; this means that 1 volt (say) is being developed at 1 Mc/s across the attenuator input; hence, if the attenuator is accurate, 10 μvolts unmodulated R.F. is appearing across the receiver Ae and E terminals. The modulation depth control may now be turned up; a common standard depth is 30 per cent, and this may be marked on the modulation depth control, or the R.F. output meter (for modulation increases the power developed) may indicate it on its scale—methods differ with various S.S.G.'s. The R.F. signal to the receiver is now being modulated, and the resultant A.F. output should be recorded on the output meter. If it is below 50 mW, the R.F. input from the S.S.G. will have to be increased, and this will indicate that the sensitivity of the receiver is below standard—or, of course, it may be found that, say, 8 μvolts is sufficient to produce an A.F. output of 50 mW, which means that the sensitivity of the receiver is above normal. That constitutes the whole of this particular form of performance measurement. It may be repeated, of course, over the whole range of frequencies handled by the receiver, and in this way any lack of sensitivity, due to poor design or defect in the receiver, may be definitely established.

Now for the "adjacent channel selectivity" test—a very important one. First, the receiver and S.S.G. are set up, as before, both tuned accurately to 1 Mc/s and the voltage injected into the receiver by the S.S.G. adjusted by the attenuator so that 50 mW output is obtained, as before. The setting of the attenuator for this is carefully noted. Next, the S.S.G. frequency is altered to 1·005 Mc/s—i.e. 5 kc/s above 1 Mc/s. It will probably be impossible to do this accurately with the main frequency control dial, so the fine control will have to be used. There are, as a matter of fact, quite a number of ways in which this 5 kc/s mistuning could be secured very accurately, but they cannot be adequately described here. Since the receiver is still tuned to 1 Mc/s, its response to this off-tune input signal will be considerably less (or should be with a modern receiver) than 50 mW output. It can be restored to 50 mW by increasing the signal strength injected by the S.S.G. by "turning up" the attenuator. The increase shown on the attenuator dial (which may, incidentally, be calibrated in terms of decibels as well as micro- and millivolts) is then a measure of the *decrease* in response of the receiver to the off-tune signal. Expressed in decibels, it may be stated, for example, as "14 db. down at 5 kc/s off-tune at 1 Mc/s."

The foregoing very sketchy outline of how to make two very important tests of receiver performance will have served its purpose if it indicates how all the parts of an S.S.G. are involved and contribute valuably to precision of information about the receiver. It is plainly impossible to do anything as exact as this with a test oscillator. Nevertheless, a well-made and carefully treated test oscillator is capable of giving practical information to experimenters and servicemen of value comparable to that offered by its more impressive cousin. It may not be possible to make absolute tests of the sensitivity of a receiver with a test oscillator, because it does not usually possess an accurately calibrated constant-impedance attenuator, nor is it usually possible, without an external valve voltmeter preceding the attenuator, to ascertain precisely the R.F. output of the oscillator. But if a number of receivers, all of the same make and type, are being tested, and the average settings of the T.O. control are found by experiment, the receiver can be classed as above or below the average by observing what T.O. control adjustments are necessary to bring the output of each one to some common standard level. The method may be made clear by considering a specific case such as a sensitivity test.

Comparative Sensitivity

First, the frequency (tuning control) of the T.O. is set at the frequency at which the test is to be carried out, and the output control is set at the average marking on its dial to give, with a receiver of the type and make under test in sound average condition, a standard output—say 50 mW, as measured by an output meter. For example, a particular make and model of domestic receiver may require the T.O. output control to be set at the 20-degree dial mark, with a 1-Mc/s signal and the T.O. internal modulator switched on. In these comparative tests, it is particularly important to specify the other factors, such as A.V.C. operative or non-operative, the nature of the load across the output stage including the output meter, the method of feeding the signal into the receiver (through, say, a 0·0005 μF condenser), and the precise voltages of h.t. and l.t. power supplies. If, now, another receiver of this make and model is tested, it may be found necessary, to obtain under the same conditions an output of 50 mW, to set the T.O. output control to the 25 deg. mark instead of the 20 deg. This would indicate, evidently, either that the sensitivity of this receiver was below average or that that of the first receiver was above average. Nor need comparative tests of this sort be confined to overall sensitivity checks. Stage gain and A.V.C. efficiency and several other aspects of receiver performance may be similarly checked.

Plainly, it would be possible—and is done, by conscientious and intelligent workers—to compile, from careful and thorough records of such tests made on a large number of receivers encountered in the course of service and repair work, a very valuable and highly practical body of information, which

would amply compensate for a lack of means to purchase more expensive and elaborate test gear. It will be obvious, however, that such comparative testing would be valuable only with three main provisos: (1) the test oscillator itself must be a soundly built and well-screened piece of apparatus of which the performance is subject to only small or negligible variations over reasonably long periods; (2) the T.O. must

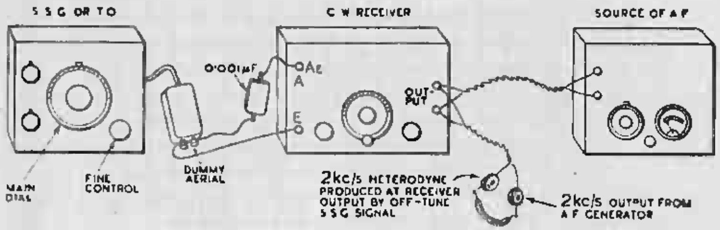


FIG. 3

An arrangement for accurately adjusting a signal to a given amount off-tune. This is essential when making adjacent-channel selectivity tests of a receiver. The C.W. receiver shown above is only used for this preliminary adjustment of the frequency of the S.S.G. output

itself be given careful treatment and periodic maintenance attention and checking of the constancy of its output and frequency calibration; (3) the conditions of test, as mentioned previously, must always be the same for any particular make and model of receiver. These are points, of course, that will be obviously necessary to intelligent operators with conceptions of the meaning of the "scientific method."

Plotting Response Curves

It may not be out of place, at the conclusion of this chapter and with the object of hinting at the many "tricks" that can be employed by skilled operators of good instruments, to describe briefly one typical method whereby, with the aid of some A.F. source covering with reasonable accuracy about 1 to 6 kc/s, an S.S.G. or a T.O. may be accurately "detuned" by a definitely known number of kilocycles on either side of some mid-frequency, for the purpose of ascertaining the "band-pass response" of a receiver, which is most important in estimating its selectivity and audio response. Assuming that this mid-frequency were 10 Mc/s, it is unlikely that the calibration of the main tuning dial of the S.S.G. or T.O. would enable accurate discrimination to be obtained within a kilocycle or two at such a high frequency. It may be easily done, however, as follows. The *un*modulated signal from the S.S.G. or T.O. is fed to a receiver capable of C.W. reception—either a "straight" type that can be made to oscillate or a superhet with a beat oscillator in the I.F. stages—and this receiver

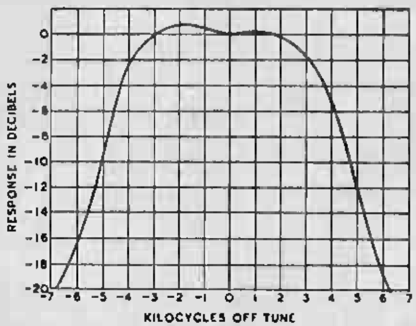


FIG. 4

Response curve obtained with the set-up shown in Fig. 3

is tuned to give zero beat in the 'phones or speaker at its output. Now across this output load is connected the source of known A.F.—at, say, 2 kc/s. The arrangement is shown in Fig. 3. Next, leaving the main tuning control of the S.S.G. or T.O. unaltered or locked, the fine control is adjusted until the S.S.G. or T.O. signal produces a beat note that coincides with that of the A.F. source, that is, at 2 kc/s. Coincidence between two A.F. tones is very easily observed with a high degree of accuracy by listening for the "warble" or very slow rise and fall (a fraction of a cycle per second) that occurs when the A.F.'s are very nearly in phase—a typical illustration of which may be taken from the "beat" between two unsynchronized engines of aircraft overhead. When the two notes produced by the arrangement in Fig. 3 are coincident as heard in the 'phones or speaker, the S.S.G. or T.O. will be detuned from the 10 Mc/s mid-frequency by exactly 2 kc/s; whether it is plus or minus depending on whether the detuning has been done in the higher or lower frequency direction. Having noted carefully the settings of the fine controls necessary to obtain the three outputs—at 9·995 Mc/s,

10 Mc/s, and 10·005 Mc/s—the sensitivity, absolute or comparative, of the receiver may be checked as previously described at these three points. The process may be repeated for 3, 4, 5, and 6 kc/s on either side of the mid-frequency, and in this way the adjacent channel selectivity of the receiver may be pictured by plotting the variations in output against kc/s off-tune in the form of a graph, as in Fig. 4, which is the familiar band-pass response curve.

It is hoped to deal with the matter of what is meant by receiver performance in modern times later on, with the object of lifting it out of the realms of guesswork in the minds of technical users of radio. But for the time being, this glimpse of the possibilities offered by instruments capable of precise quantitative measurements will have to suffice.

CHAPTER III

OUTPUT METERS AND ATTENUATORS

FIRST, it is necessary to make clear the meaning of the term "output." In the case of a receiver for sound reproduction, or of an A.F. amplifier, the output is conventionally taken to be the rate at which electrical energy is imparted to the speech coil of the loudspeaker or the windings of headphones. In other words it is an electrical power output which may be expressed in watts or milliwatts and measured by a form of a.c. wattmeter.

This way of estimating the output of a receiver or amplifier is simply a convenient convention. The true output consists, after all, of energy imparted to air by the motion of the diaphragm which is attached to the speech coil. This true output manifests itself in two ways: by physical movements of air particles and by heating effects. The reason why quantitative measurements of these two phenomena are not usually made or attempted, save for special purposes, is that they are difficult to carry out accurately and require rather expensive and elaborate apparatus. The human ear itself is, as a matter of fact, one form of "meter" responding to the true output. But it is incapable of precise and objective measurement, and serious errors are introduced, even with the best trained ears, by environmental conditions. It is possible to make measurements of the true sound output by systems of apparatus such as calibrated microphones and associated gear, with the speaker suspended in a space free of echo and acoustic resonance effects, but even then it is not easy to obtain anything but comparative data—and for the great majority of practical, everyday purposes data obtained under these special conditions are of little use because of the extremely variable and uncontrollable conditions under which speakers are normally employed. However, as a matter of interest, it may be mentioned that a few years ago *Wireless World* inaugurated routine tests on these lines and published curves of various makes and models of loudspeaker.

We have normally, then, to take for granted the efficiency and fidelity with which a speaker converts the electrical output of a receiver or amplifier into sound; performance measurement

normally stops at the electrical output fed to the speech coil.

Now there are always at least three factors that have to be considered or understood when describing a.c. power phenomena. These are (*a*) the waveform (fundamental and harmonic content); (*b*) the magnitude in volts or amperes; (*c*) the nature and magnitude of the impedance in which the power is being dissipated or used up. If these factors in the test input applied to a receiver or amplifier are accurately known, it is possible, by comparing the resultant measured output against this input, to estimate the amount of gain, distortion, etc., brought about by the operation of the circuits between the input and output terminals.

Test Routine

To make matters as simple as possible it is usual to employ a pure sine-wave voltage input of known amplitude when measuring performance involving output measurements. This is usually produced by a beat frequency oscillator, of which the frequency can be varied from 20 to 15,000 c/s; the amplitude of this voltage is monitored, so that it is kept always the same, by a valve voltmeter. To enable accurately known fractions of this constant voltage to be developed across the amplifier input terminals as may be required, an attenuator is used; and an output wattmeter records the electrical power fed to the speaker or dissipated in an impedance of equivalent magnitude.

The matter of the harmonic content of the output has to be dealt with by rather more complex gear. "Harmonic distortion" may perhaps be explained in very simple terms by saying that it consists essentially of the appearance in the output of frequencies that were not present in the input. They must have been put there somehow by the circuits handling the test signal.

There are two ways at least in which this sort of distortion can not only be made manifest (apart from the often unpleasant effect on the ear) but actually be measured. One way is to inject the output into a cathode-ray oscilloscope so that an actual picture of the waveform appears on the screen; the picture can be copied accurately on to paper and analysed by mathematical process (Fourier's analysis) into the component sine waves that make it up, the height and frequency of these component waves being thus ascertained and so the proportion they form of the total output. Another way lies in setting up sharply and accurately tuned filter circuits that separate the component waves and enable each one to be examined for amplitude and frequency on its own; these are called "harmonic filters" and are often quite complex, bulky, and expensive pieces of apparatus. The oscilloscope is a very convenient modern instrument for ascertaining the harmonic content of the output of a receiver or amplifier, and is very commonly used for the purpose; harmonic filters are far less commonly used, and then seldom outside a laboratory.

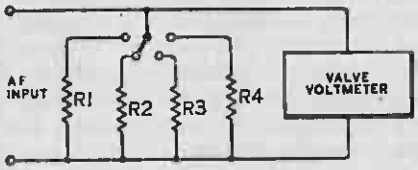


FIG. 5

A simple output meter. The resistances R_1, R_2, R_3, R_4 may be of ohmic values corresponding to the impedances of the various normal loads in the anode circuits of output stages, e.g. 2Ω and 8Ω to represent speech coils; 2000Ω and 12,000Ω to represent primaries of speaker transformer. The valve voltmeter must be able to cover the ranges of voltage developed across these resistances

Now the output meter forms part of the impedance in which the output power is dissipated. It is necessary, therefore, if reliable results are to be obtained, that the output meter should either have so high an impedance that its shunting effect across the load impedance is negligible, or that it should provide, in conjunction with the load impedance, the correct output test impedance. Briefly, therefore, output meters may be classified into two types. In the first type to be considered the actual meter is a valve voltmeter, and this measures or records the voltage developed across a fixed known resistive output impedance that dissipates the output power in the form of heat; or it may consist of a low-impedance a.c. ammeter in series with the load impedance, measuring the current through that impedance.

The general principles involved are as follows. The power W, dissipated in the load resistance R, is given respectively by

either $W = I^2R$ or $W = \frac{V^2}{R}$. Plainly, the value of R in ohms must be equal to the reactance in terms of ohms of the normal load impedance—the speaker speech coil, for instance—at the frequency employed during measurement. For this reason a multi-range output meter of this type may consist of a number of load resistances selected by a switch, across which, or in series with which, a valve voltmeter or an a.c. ammeter respectively may be connected on an appropriate voltage or current range. For example, a quite useful form of output meter may be constructed on the lines of the circuit shown in Fig. 5, and the dial of the meter may be calibrated in terms of watts instead of in volts. Such meters, unfortunately, are not very sensitive and are not used much for measurements of lower power output of the order of less than a few milliwatts.

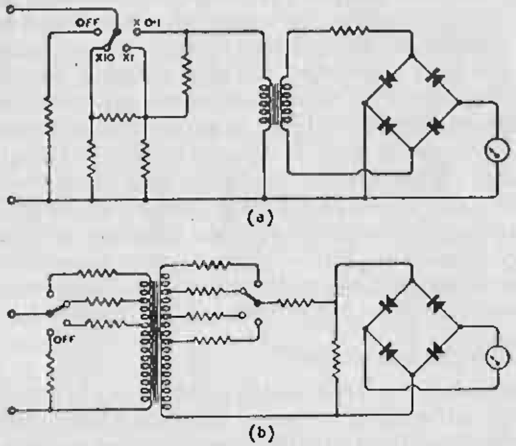


FIG. 6

In (*a*), the impedance of the instrument is constant, but the range of power measurement is variable. In (*b*) both the range of measurement and the impedance, for matching purposes, are adjustable. Note the "Off" switch position; a powerful output stage should always be loaded to avoid damage. No d.c. should be passed through the transformer windings

The second type of meter to be considered makes use of the usual metal rectifier type of voltmeter and a transformer, the inductance of the primary of which makes it less independent of frequency than the former type. But it can be made sufficiently sensitive to measure outputs of the order of microwatts.

Two practical forms are shown in Fig. 6 (*a*) and (*b*), which give skeleton circuits. If a number of receivers of the same type, requiring the same output load impedance, are to be tested, the impedance offered by the output meter and load resistance may be kept fixed, and various output ranges measured by making the load a resistive network offering approximately constant impedance at all settings of the range switch. Or, in a more versatile type, the input impedance as well as the range of the output meter may be altered to suit various receivers by tappings on the transformer primary and secondary.

For the sake of completeness it should be stated that the meter reading in rectifier-type instruments is proportional to the average value of the input waveform, and that errors may occur if the meter is calibrated with a sine wave and used subsequently on waveforms with a high harmonic content.

Comparative Measurements

The foregoing is a brief sketch of the sort of meter that can be used to make fairly accurate absolute power output measurements. These are usually expensive instruments, and for a great many purposes in the general run of service and test work only comparative data are required. For example, in an amplifier it is usually not so much the actual power output that is interesting, but the levelness of the response of that amplifier to a wide range of frequencies. Thus, it is more useful to know, very often, that the output power for a constant input voltage is much the same at 100 c/s as it is at 2000 c/s—i.e. these two outputs have to be compared, and not exactly measured.

For such purposes the a.c. ranges of many types of general-service multi-range meters are quite suitable, and many of them are actually designed to be used in this way, as is shown in Fig. 7, which shows how an output meter may actually be connected to, say, the output stage of a domestic receiver for such purposes as observing resonance when lining up I.F. stages or taking audio frequency response tests and so on. It must not be forgotten that a powerful output stage must never be operated on "no load," i.e. with the speech-coil circuit completely open-circuited, otherwise peak voltages, especially at high A.F., may seriously damage insulation.

Either the output meter (representing the speech coil) *or* the speech coil itself, must always be in circuit to provide a load.

A very simple arrangement is given in Fig. 8 of a 2-volt battery valve and an ordinary d.c. 5 mA meter with a few load resistances that may be used successfully for quite a number of comparative indications. This inexpensive little gadget is handy when aligning the R.F. and I.F. stages of a superhet receiver and even when taking rough frequency response curves in amplifiers. Since only comparative indications are noted, the dial of the meter is not altered from its original scale. A triode valve of the H.F. or "Det." type will pass about 5 mA with applied a.c. of from 10 to 15 volts, and does not seriously shunt a low load resistance.

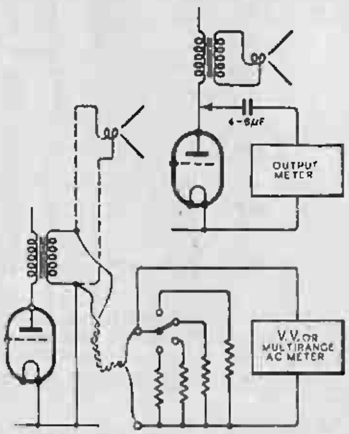


Fig. 7

The speech coil of the speaker may be left in cicuit or replaced, as shown, by an equivalent resistance. If direct connection to the anode of the output valve is needed, it should be made through a good high-capacity paper condenser, not an electrolytic

Attenuators

Attenuators, although not "instruments" in the true sense of the term, are accessories of such value that they form a very important part of the outfit of a serious experimenter or tester. Their study covers a very wide and abstruse field, and in this chapter it is necessary to narrow it down to that portion containing the types of a.c. voltage attenuator of immediate practical use for R.F. or A.F. work.

The reason for their use and the source of their value is best understood by considering some practical example. How, for instance, is "one microvolt" measured, especially when it is at, say, 10 Mc/s? The answer is that it is not measured; it is "picked off," as it were, from a source developing an accurately measured 1 volt. The 1 volt across, say, 10 ohms may be quite easily and accurately measured by some form of R.F. meter, either a valve voltmeter or a series thermo-couple arrangement.

The business of the attenuator is then to provide, across its output terminals, some known fraction of this original 1 volt.

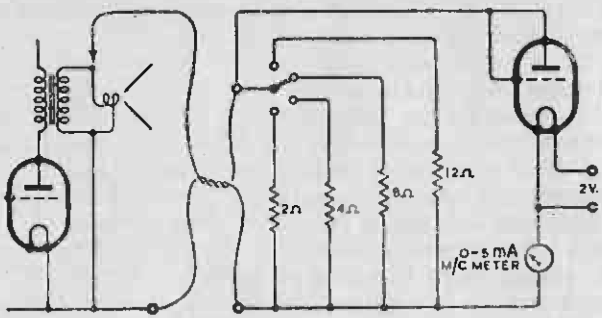


FIG. 8

For comparative indications it is seldom necessary even to disconnect the speech coil, the arrangement being as shown. The choice of load resistance is determined by the maker's figures for speech coil impedance, and/or the ratio of the speaker transformer

Were it d.c. that was being dealt with, the matter would be quite simple, but a.c. involves a number of factors that add considerably to the complexities and difficulties of design of attenuators.

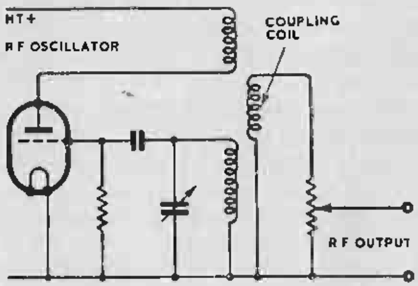


FIG. 9

This type of output control is used in several inexpensive test oscillators, and is satisfactory for rough general service work

The "attenuator" of a simple and inexpensive service test oscillator—very rightly more often termed merely an "output control"—may consist of a simple potential divider across a low impedance coupling coil energized by the valve oscillator

as represented by Fig. 9. The advantages of such an arrangement consist mainly of simplicity and cheapness, and for a good deal of general service work it is quite satisfactory. But it is practically useless for accurate quantitative measurements. It certainly "controls" the magnitude of the R.F.

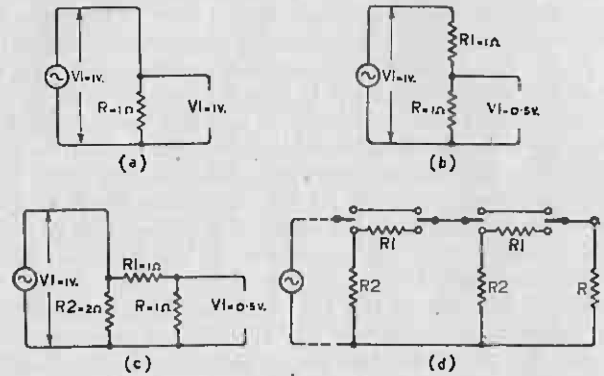


FIG. 10

Development of the "inverted L" attenuator. By means of the switches shown, any desired amount of attenuation can be obtained in steps. The dotted line indicates where other "L" cells may be added

voltage appearing across the output terminals of the test oscillator, but not only is the absolute value of this voltage not known, but, unless the input impedance of the apparatus under test which is connected across the oscillator output terminals is extremely high, and so constitutes a negligible load, the frequency of the test oscillator is altered with every movement of the output control slider, owing to the "reflection" of the load across the coupling coil back into the tuned circuits of the oscillator.

What is wanted is something that provides accurately known attenuation of the voltage across the coupling coil *and* presents, all the while it does so, a *constant* load on the oscillator tuned circuits, so that neither the frequency nor the magnitude of the voltage across the input terminals of the attenuator vary greatly with the amount of attenuation. This is commonly done by means of a "constant impedance R.F. voltage attenuator," which is not a very simple, and seldom an inexpensive, piece of apparatus.

The theory of such attenuators is dealt with in several books; the design problems are greatly simplified by standard formulae for various types.

One of the simplest and most effective types is the "inverted L," the theory of which may be explained as follows. Referring to Fig. 10 (*a*), it will be assumed that 1 volt at 1 Mc/s is being developed across R, the terminating impedance (1 ohm). In Fig. 10 (*b*), R is in series with another resistance R_1 of 1 ohm, so that the voltage across R is now 0·5 volt, or half V, or "6 db. down." But this, obviously, is not now presenting 1 ohm impedance towards the generator (valve oscillator), so that it is likely that the frequency of the oscillator will alter. However, if, as shown in Fig. 10 (*c*), another resistance R_2 of 2 ohms is placed in parallel with R and R_1, the attenuation of – 6 db. is still obtained, but the impedance of the network towards the oscillator is still 1 ohm as it was in Fig. 10 (*a*). This process, of adding "L" cells to the left of R, can be continued indefinitely, each cell adding – 6 db. attenuation without altering the load on, and therefore the frequency of, the oscillator. By calculation of the values of the "series" and the "parallel" elements of the "L" cell, attenuation in steps of any required number of db. can be secured.

It is common to construct attenuators to work in 10 db. steps, to facilitate measurements. Thus, such an attenuator with 8 steps, each of 10 db., would provide a voltage across the terminating impedance (R) – 80 db. down, or 1/100,000 of the voltage developed across the input terminals of the attenuator. Hence, if V were 1 volt, 10 microvolts would be obtained.

A quite common form of attenuator used in standard signal generators of the general-purpose type is that shown in Fig. 11. Here, the "coarse" control of output is effected by a network that attenuates the voltage by 10 db. steps. For intermediate values, the voltage across the input terminals of this "multiplier" section is continuously varied by a plain slider arrangement. This is called a "ladder" attenuator, and is obviously not of constant impedance looked at from the oscillator. However, the ratio of change of attenuation to the change of load on the oscillator at various settings of the controls may be made high enough for any changes in frequency and departure from attenuator calibration figures to be practically negligible. Thus, such an attenuator may increase

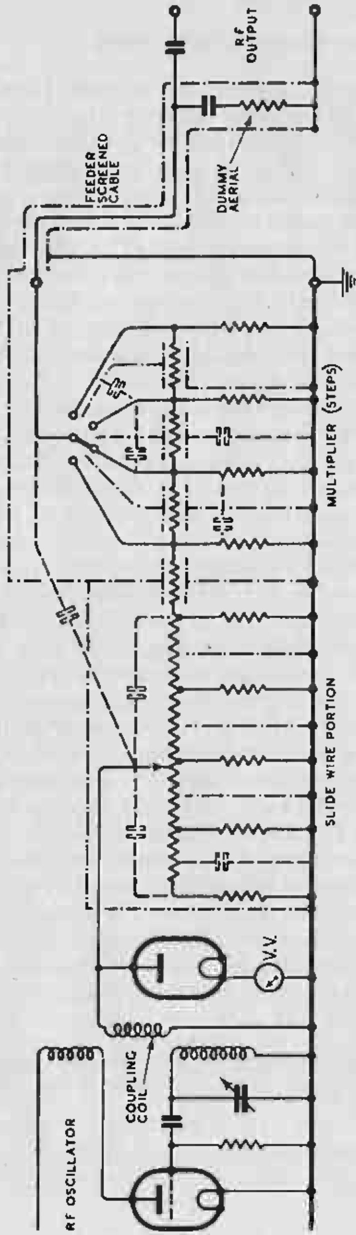


FIG. 11

Standard signal generator attenuator. The input voltage is monitored by a diode valve voltmeter. This voltage can be attenuated in steps by the "multiplier" and continuously varied in the "slide wire" potential divider. Some possible stray capacities are indicated, and also the elaborate screening required between elements or decrease the voltage across the output terminals by as much as 100,000 times while making the corresponding increase or decrease of load provided by the attenuator on the oscillator only double or half, even at extreme ranges of attenuation. By low impedance loose coupling between the attenuator and the oscillator tuned circuits, the effect on frequency of this change of load may be made negligible. The calculation of the values of the resistive components of this type of attenuator is not very difficult;* it starts with the value of the terminating resistance and the amount of attenuation per step required. Since this terminating resistance has to be correct if the calibration of the attenuator is to be valid, standard and similar precision signal generators are normally provided with standard "dummy aerials" and standard feeder cables, which constitute this terminating impedance, and it is necessary that the circuit under test should be of high enough impedance to constitute a negligible shunt across this dummy aerial and feeder.

The construction of R.F. voltage attenuators, especially at the really high frequencies, is rendered difficult and complex by the presence of stray capacities in addition to the resistance network, as shown in dotted line in Fig. 11. These cannot be completely eliminated, but they can be made small and their effects allowed for by special mechanical constructions. For example, a well-known form of R.F. voltage attenuator consists of a heavy brass casting containing compartments for each resistance and, perhaps, moulded slider and switch parts; it is easier and cheaper to have a casting than to build up a complicated structure from sheet and tube metal. There are, of course, several other types of attenuator, but the ones thus briefly described are those most generally used in quantitative measurements on ordinary receivers. Their construction is comparatively simple, they can be designed to hold their calibration up to all but ultra-high frequencies, and they are mechanically well able to stand up to the hard use of routine testing.

* See *Radio Laboratory Handbook*, by M. G. Scroggie, *Measurements in Radio Engineering*, by F. E. Terman, *Wireless World* "Radio Data Charts," by R. T. Beatty.

CHAPTER IV

VALVE VOLTMETERS

The special property of the valve voltmeter is its very high input impedance. That means that the reactance or resistance existing across its test terminals, due to stray capacities, resistances in the circuit of the valve voltmeter, the effects of any valves employed, and such factors, is extremely great. This is a very desirable quality in any voltmeter, since its connection to the circuit under test should make the least possible difference to the working conditions being examined. It must, however, be remembered that this excellent property of the valve voltmeter is obtained at the expense of cheapness, strength, reliability, simplicity, and freedom from the necessity for power supplies, which are usually the benefits enjoyed by the use of well-made rectifier-type meters. Nor is the valve voltmeter very sensitive with low inputs, save in special laboratory forms unsuitable for general use. But the high input impedance of the valve voltmeter, especially when dealing with radio-frequency circuits, makes it almost indispensable in certain kinds of measurement. If a valve voltmeter does not possess a high input impedance, or if the circuit under test is of low frequency and impedance itself, there is no point in using a valve voltmeter at all; it comes into its own only for measurements in circuits of high impedance themselves involving frequencies so high that even small capacities introduced by the measuring instrument would act as an appreciable shunt and seriously upset the working conditions.

It may be best, perhaps, to begin by examining the skeleton circuit of a very commonly used valve voltmeter, such as is given in Fig. 12. Although it looks, to a novice, a rather complicated arrangement, it may actually be divided for the purposes of study into five main sections, each of which is in itself quite simple to understand. These five sections may consist—referring to Fig. 12—of (*a*) a diode rectifier with a high value of load resistance; (*b*) a triode to the grid of which the voltage developed across the load resistance is applied, thus altering the conductivity of this triode; (*c*) a "bridge" circuit, including the triode as part of the network, of which the function is to prevent the indicating meter having to carry the anode current of the triode, thus enabling a more sensitive meter to be used for indicating the unbalancing of the bridge when the conductivity of the triode changes; (*d*) a cathode resistance variable in steps that, in a way to be explained in due course, changes the range of measurement; (*e*) a potential

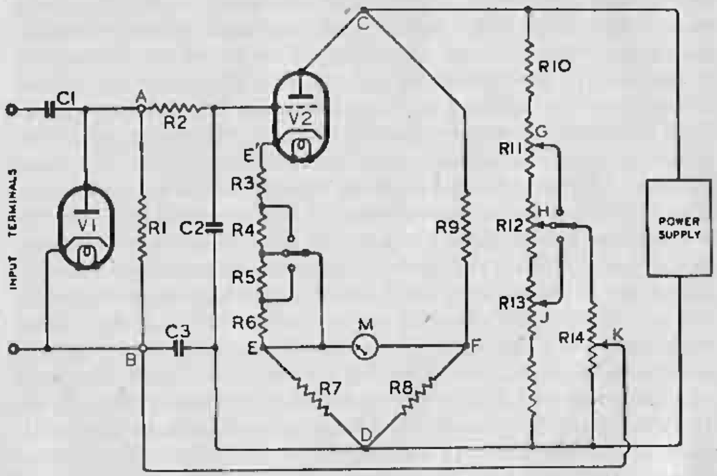


FIG. 12
Skeleton circuit of a commonly used form of general-purpose valve voltmeter

See Scroggie Sect 69

divider across h.t. positive and negative that provides a potential for the triode grid enabling it to operate in this circuit over the straight part of its anode current–grid voltage characteristic, giving linearity of scale indication throughout the range and maximum sensitivity.

These five sections are shown separately for reference purposes in Fig. 13, and slight alterations have been made in order that their individual actions may not be confused. When the behaviour of each section is studied individually and its function is understood, their action in combination will be made clear.

Rectification

Fig. 13 (*a*) shows the diode rectifier and its load, R_1, and the feed condenser C_1. With no alternating voltage across the input terminals, a very minute potential difference, if any at

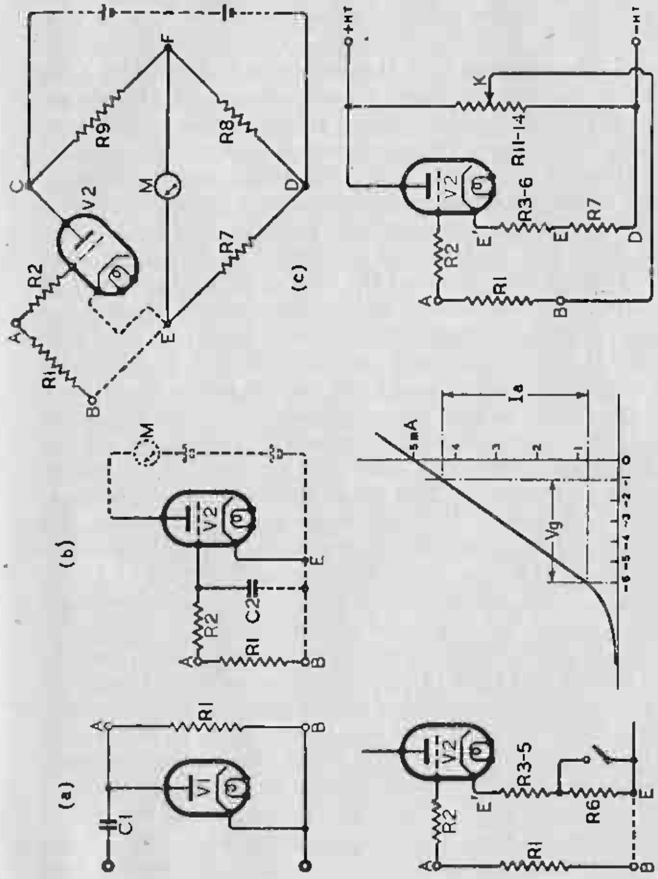


FIG. 13
Five separate functions are performed by the circuit of Fig. ~~11~~ 12, which is here dissected into its constituent parts

all, exists across R_1. If an alternating voltage is developed across the input terminals, however (it will be assumed that it is of sine wave form), the diode will pass current during the positive half-cycles only. Now, an easy way of remembering the sign of the voltage polarity in any electrical circuit is to say "P, positive, poor in electrons—N, negative, numerous electrons." The anode side of the condenser C_1 begins by being neutral—i.e. having its correct complement of electrons—but when the positive half-cycles of voltage occur on the input side of the condenser, the electrons in the anode side are strained towards the input side; the diode anode, which is connected to the anode side, thus acquires a positive potential in respect to cathode, during the positive half-cycles, and electrons from the cathode are attracted to it during these periods. Having once reached the anode, they cannot get off it again (because the anode, unlike the cathode, is not hot and emissive) save through the resistance R_1. If R_1 is of high value—and it is normally made anything from 5 to 100 MΩ—they do not get through it easily and remain crowded—numerous—on the anode side of the condenser C_1. Thus the anode acquires, with each succeeding positive half-cycle, an increasing *negative* potential; this goes on until the negative potential so developed at this point *equals the peak voltage of the positive half-cycles.* If this voltage can be measured, the R.M.S. value of the alternating sine-wave voltage across the input terminals AB can be derived by multiplying by 0·707.

Now, since R_1 is a comparatively high resistance, no ordinary moving-coil and resistance type of voltmeter will be able to indicate the potential difference across it satisfactorily, because even the best of this type of voltmeter has an input resistance of only a few score thousands of ohms, which would act practically as a short circuit across R_1. If, however, R_1 is made to act as the "grid-leak" of a triode valve, the voltage across it can be made to affect the anode current of this triode valve in an easily observed way. This is shown in Fig. 13 (*b*), where the differences between this arrangement and Fig. 12 are shown in dotted line. If the triode valve starts by passing, say, 5 mA when there is no voltage across R_1, the development of a voltage across R_1, by the action of an alternating test voltage applied across the valve voltmeter input terminals to the diode as explained above, may cause the anode current of V_2 to change to 1 mA, which is readily seen on an ordinary meter. Since the voltage at the diode anode is negative, no grid current will pass through R_1, and hence no power will be used, and the only resistance across R_1 will be that of the insulation between V_2 grid and cathode, which will be of the order, normally, of thousands of megohms. If the grid of V_2 were connected directly to the diode anode, the grid-cathode and grid-anode capacities and the stray capacities of the wiring might, at really high radio frequencies, enormously reduce the input impedance of the valve voltmeter—even 10 $\mu\mu$F offers reactance of only 1600Ω at a frequency of 10 M/cs. Hence, R_2 is included to provide adequate impedance to the alternating voltage input that is to be measured. The grid of V_2 may then be safely rid of any R.F. by the by-pass condenser C_2. There is no voltage drop through R_2 because no grid current is being passed; what gets to the grid of V_2 is the full voltage developed across the resistance R_1.

It is easy to observe, on an ordinary meter, a comparatively large change in the I_a of V_2. If, however, small changes, caused by small test inputs, must be read easily, this meter must be a sensitive one—say a microammeter. But such a meter cannot carry the standing I_a of V_2, which may be of the order of 5 mA. For this reason, a bridge circuit is employed, as shown in Fig. 13 (*c*). In this V_2 forms one of the resistive paths in the bridge network, and the meter is used only to indicate when the bridge is out of balance. At other times it remains at zero. R_7, R_8, and R_9 are of such values that when V_2 is passing 4·5 mA (which will be assumed to be when its grid is 1 volt negative with respect to the cathode of V_2) it acts as a resistance that completes the exact balance of the bridge. Now if a voltage is developed across R_1, the conductivity of the valve is altered and the bridge is thrown out of balance. The sensitive meter M indicates readily the slightest unbalance of this bridge, and if its scale is suitably calibrated, the precise amount of the unbalance, which in turn depends on the voltage across R_1, which is equal and opposite in polarity to the peak of the positive half-cycles of test voltage. Hence the meter M may be calibrated for sine-wave inputs in terms of R.M.S. values.

Referring to the I_a/V_g curve shown in Fig. 13 (*d*), it will be seen that the amount by which the grid potential can change toward negative with respect to the valve cathode is strictly limited, in this case to about − 6 volts; beyond that, it runs into the curved part of the characteristic, which renders scale indications no longer linear—i.e. the divisions can no longer be evenly spaced—and, further still, it runs to "cut-off" and no indications are obtained at all. It is theoretically possible, of course, to measure up to quite high test voltages by taking bits of it at a time, say every 6 volts, but this is an impracticable and unnecessary method.

From Fig. 13 (*d*), it will be seen that resistance is included in the cathode circuit, i.e. R_3–R_5 and R_6. If V_2 passes 4·5 mA of I_a, the passage of this current through the cathode resistances sets up a voltage across them. This is well known as the principle of "automatic bias" in domestic receivers. The cathode, E', will be positive with respect to E. Hence, if R_1 is connected to E, as shown for the sake of argument in the dotted line, the potential of the grid will be negative with respect to E', or, E' will be positive with respect to the grid by an amount depending on the I_a or V_2 and the ohmic value of the cathode resistances. If a voltage is developed across R_1, then the I_a of V_2 will change—and so, therefore, will the voltage across the cathode resistances. If the point A becomes negative, V_2 will pass less I_a. Hence, E' will become less positive with respect to E, and therefore with respect to the grid of V_2. Which is much the same as saying that instead of the grid becoming negative with respect to E' by the full amount of voltage across R_1, which would happen if the voltage across the cathode resistances remained constant, it becomes negative by some amount *less* than the full voltage across R_1. Plainly, the larger the ohmic value of the cathode resistance, the greater is this effect. If the cathode resistances are made, say, of the order of many thousands of ohms, the voltage across R_1 can become quite large before the *net* change in the grid potential of V_2 with respect to the cathode E' of V_2 is such as to extend beyond the lower end of the straight part of the I_a/V_g curve. This means that the test voltage input to the valve voltmeter can be measured over a wide range in one sweep, and changes in the range of measurement of which the valve voltmeter is capable can be made very simply by altering the value of the cathode resistance. In this case, referring to Fig. 12, it is done by switches and a chain of resistances.

Range Switching

The last portion of the circuit to be studied presents no difficulties if the foregoing has been understood. It is plain that if the "earthy" end of R_1, the point B, were taken to E, the effective potential of the grid of V_2 would settle down to some value dependent on the magnitude of the cathode resistance in circuit, and would therefore change with each movement of the range switch; the I_a of V_2 would also change, and the bridge would have to be balanced by altering the ratio of R_9 to R_8 each time. Again, if B were connected to D this point is so far negative with respect to E' that very little I_a would be passed by V_2 and again the bridge would be difficult to balance save by considerable alterations in the bridge elements. Hence, the point B is connected to the slider of a potential divider across h.t. positive and negative. If only one range of test voltage were to be measured, the slider could be set at some one point only. But since several ranges are involved, each demanding a different initial potential for the grid of V_2 in respect to cathode, the potential divider is made up of a chain of potentiometers, and the correct working potentials selected by a switch ganged to the range switch. Final adjustment is made by a slider on a main potential divider, R_{14}. The potential of the point K is made (in this case) just one volt less positive in respect to h.t. negative than E', which means that the steady (no input) potential of the grid of V_2 will be $-$ 1 volt with respect to E'; and V_2 will pass the current required for balancing the bridge.

Finally, mention may be made of the means whereby the input impedance is kept high even at high radio frequencies. The diode V_1 is usually an acorn type and is carried in a low-loss holder on which are mounted also R_1, R_2, and C_1. This holder or "probe" is connected to the triode and bridge circuits and the power supply, which are all mounted in a separate casing, by a flexible cable that contains the leads from the cathode of V_1, the heater or filament leads, and the lead from the grid (V_2) end of R_2. By this means, the diode anode feed condenser can be brought right up to the circuit being tested and leads to it kept as short as possible, which reduces the capacity across the input terminals and the inductances of these leads to the minimum. The calibration of such a well-constructed valve voltmeter is extremely simple. Since the input impedance is so high at all frequencies, it may be calibrated with 50 c/s mains supplies, the voltage being reduced by suitable combinations of step-down transformers or resistive potential dividing networks, with an ordinary metal-rectifier type of a.c. meter as the voltage standard. Calibration of the valve voltmeter obtained thus will hold good at frequencies up to the order of 20 or 30 Mc/s and even beyond with reasonably small errors if used with care. There are several other kinds of valve voltmeter, of course, both commercial and laboratory, but if the one here described is clearly understood, little difficulty should be experienced in grasping the principles of the other types.

CHAPTER V

TESTERS AND BRIDGES AT A.F. FOR INDUCTANCE AND CAPACITANCE

It should be recalled that inductance and capacitance are *properties* of electrical circuits. Inductance can be deliberately conferred on a conductor to a given amount by making it of a calculated shape and size. Capacitance can similarly be brought into being by special arrangements of conductors in space. From this it follows that fundamentally it is possible to calculate, from the shape, dimensions, and positions in space of conductors and a knowledge of the electrical properties of the medium surrounding them, the amount of inductance or capacitance present. There are, in fact, many formulae enabling these purely physical and mechanical factors to be estimated with some precision.

However, it is not often convenient and easy to estimate the amount of inductance or capacitance present from physical data. It is commonly easier to observe the effects produced in electrical circuits and deduce from the measurements of these effects how much inductance or capacitance is present.

Perhaps the prime effect of both inductance and capacitance is that of reactance to a.c. This corresponds in a.c. circuits to resistance in d.c. circuits (up to a point, of course) and can be shown in a very easy way simply by including the inductance or capacitance in a circuit carrying a.c. and observing what difference to the current flowing is made by the presence or absence of the inductance or capacitance. The circuit is basically that shown in Fig. 14 (*a*). The similarity between this circuit and that of a circuit suitable for measuring resistance with d.c. will be apparent if it is imagined that the alternator V in (*a*) is replaced by a battery, and Z by a known variable resistance. The procedure would be to adjust the variable resistance with terminals 1 and 2 short-circuited until a certain value of current, measured by the ammeter, were passed. Then the short-circuit connection between 1 and 2 would be removed, leaving the unknown resistance, corresponding to Z_x, in circuit. The addition of this resistance to the circuit would cause a fall in the current passed, depending on the value of the unknown resistance, which could be readily calculated by simple Ohm's Law formulæ, from a know-

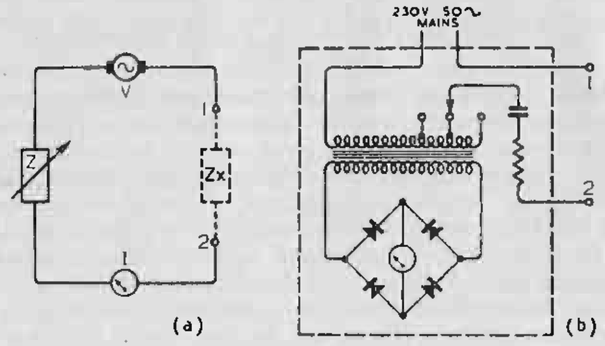


FIG. 14

(*a*) Resistance or impedance measuring circuit in which the addition of the unknown element Z_x makes a change in the current flowing in the circuit. (*b*) Simplified form of circuit with provision for reactance measurement. A series resistance is included in the primary circuit to prevent resonance, and the capacity is to stop d.c.

ledge of the value of the variable resistance, and of the voltage of the battery. This is, of course, the sort of circuit that is widely used in the resistance-measuring ranges of ordinary multi-range meters using an internal battery and adjustable resistance, the scale being marked in terms of the unknown resistance connected across the test terminals.

The same thing can be done using a.c. mains as the source of e.m.f., as shown in simplified form in Fig. 14 (*b*), and arranging the impedance of the primary circuit of the current transformer in the rectifier type a.c. ammeter commonly used so that with the test terminals 1 and 2 short-circuited, the pointer of the meter reads at full-scale deflection. When the short-circuit across the terminals is replaced by the unknown impedance, the current is not so great, and the pointer falls to a degree marked in terms of the unknown impedance.

Snags exist in this apparently very simple arrangement. The word "impedance," rather than either inductance or capacitance, has been purposely used, because, especially in the case of an inductance such as a choke or similar A.F.

winding, the component or circuit being tested very often contains appreciable resistance as well as reactance or may consist of mixed inductive and capacitative reactance as well as resistance, the whole forming an impedance. Now, the effect of a mixed impedance in a circuit is very different from pure reactance. In Fig. 15 is shown what may actually be encountered in practice. Fig. 15 (*a*) shows an unknown condenser to be measured placed across the test terminals of a

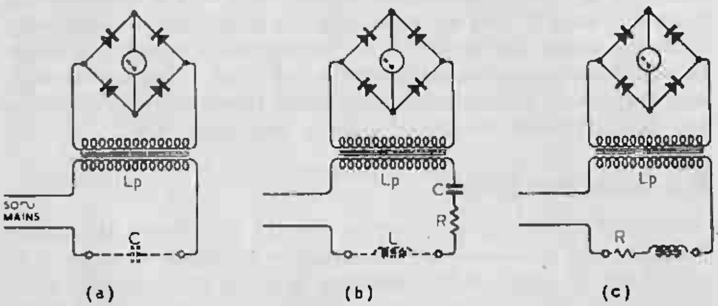


FIG. 15

(*a*) A possible source of error in this circuit is that L_p and the test condenser *C* may form a series resonant circuit at the a.c. mains frequency. (*b*) A similar danger exists in this circuit if inductance is connected across the test terminals. (*c*) Resistance in the inductance windings may, if it forms a considerable proportion of the total impedance, invalidate the calibration of the inductance measurements in an instrument designed to make them

multi-range meter provided with a scale calibrated in microfarads for such measurements using a.c. mains. The makers of commercial instruments of this kind take good care to avoid it, but it is not impossible that home-made instruments may be so constructed that the inductance of the current transformer primary L_p in combination with the capacitance of the unknown test condenser *C* form a series *LC* circuit tuned to the fundamental of a low harmonic of the mains supply. This unfortunate coincidence will manifest itself in the destruction of the meter or the blowing of the main fuses, since such a circuit offers extremely low impedance to a.c. of the resonant frequency.

In commercial instruments, therefore, it is usual to include so much resistance in the primary circuit, together with sufficient fixed capacitance, as to render it substantially non-resonant to mains frequencies no matter what unknown capacitance is connected across the test terminals. Fig. 15 (*b*) shows what may exist if the component being tested happens to be a choke connected in a circuit that is intended only to measure capacitance. Although the addition of any amount of capacitance might not render the circuit resonant to mains frequencies, the addition of inductance may—with disastrous results. $L_p + L$ may form with C a series resonant circuit, the whole of the power being carried effectively by R and burning it out. C and R may be capacitance and resistance purposely included inside the meter on the capacitance measuring range to avoid resonance under proper conditions. Needless to say, any approach to resonant conditions completely invalidates the calibration of the meter scale in any case.

D.C. Resistance Errors

A third possibility is shown in Fig. 15 (*c*). There the choke being tested has such high resistance windings that its d.c. resistance R forms a considerable part of the impedance it offers. This is quite often the case. The impedance of L having a d.c. resistance R_L is given by $\sqrt{\omega L^2 + R_L{}^2}$. If R_L is considerable, as happens with thin-wire windings in small size chokes, the impedance in ohms is much more than the reactance offered by the inductance of the choke alone, and the meter scale calibration offers very unreliable indications in terms of inductance alone.

These simple methods of measuring inductance and capacitance by applications of "Ohm's Law for a.c. circuits" have, evidently, serious limitations. The difficulty and expense of providing adequate a.c. supplies at the power and voltage levels required, and the comparatively high audio frequencies that would be needed to give reliable indications with unknown inductances and capacitances of small values, lead to most of the "universal" types of multi-range meter employing a.c. mains at 50 c/s and over 200 volts as the supply. This low frequency severely curtails the limits of the inductance and capacitance values that can be reliably measured on them. Usually they are only fit to measure inductances between about 0·5 to 50 H, and even then they must be of fairly high Q (i.e. have low d.c. resistance windings). The equivalent capacitance range is from about 0·1 μF to some 10 μF. Electrolytic

condensers cannot be tested on them, nor do they indicate the power factors of condensers and coils. However, for general A.F. work, they are very useful, since low values of inductance and capacity are not often involved and the accuracy of measurement need not be very high.

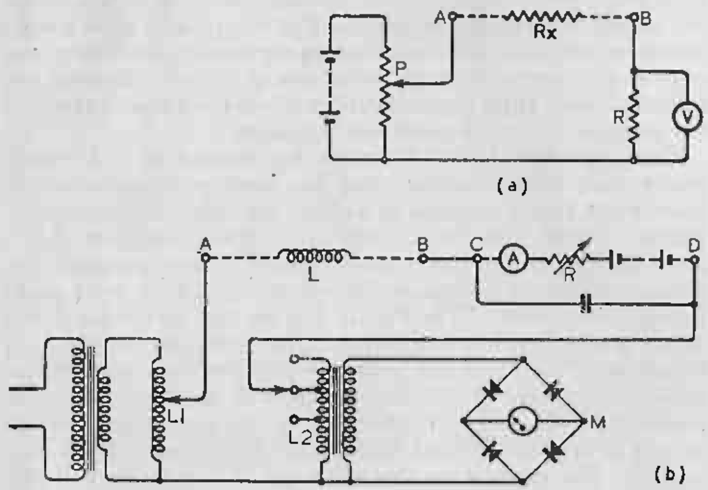


FIG. 16

(*a*) Illustrating the basic principle of a "constant voltage or current" resistance measuring instrument. (*b*) The same principle is employed in the a.c. version for measuring inductance, and the accessory terminals *C–D* enable known current to be passed through the windings and the effect on inductance to be estimated

A rather more serious limitation lies in the difficulty, in the case of inductances with iron or alloy cores, of observing the changes in effective inductance with the passage of various amounts of d.c. through the windings. This is often of great importance. Unless it is to be used in a circuit carrying pure a.c. of limited peak value, knowledge of the open-circuit inductance of a choke or transformer winding is liable to be very misleading. For example, the inductance of the primary of an output transformer may be, when measured with no d.c. flowing through the windings, as high as 40 henrys. But with the anode current of the output valve passing through it, say 30 mA, it may fall to 7 or 8 henrys, and any calculations about output and fidelity made on the basis of a 40-henry inductance as the valve load be rendered meaningless. Again, a small "midget" intervalve transformer with a "high-mu" alloy core may possess an astonishingly high open-circuit inductance—which practically vanishes when even a few mA are passed through the windings. For this reason, good manufacturers of reliable A.F. inductances are usually careful to state, or even give curves showing variations of, with d.c. flowing, the inductances of their products under given conditions involving the passage of d.c. through the windings.

One ingenious form of circuit for measuring inductance under such circumstances, that has been commercialized in convenient forms, consists of a slight variation of the application of "Ohm's Law for a.c. circuits." The basic form of it is given in Fig. 16 (*b*), but a consideration of an equivalent d.c. circuit employing resistance, shown in Fig. 16 (*a*), may make its operation clear. If, in Fig. 16 (*a*), the test terminals A and B are short-circuited, the slider of the potential divider may be adjusted to a position near the negative end so that just sufficient voltage is developed across R to bring the voltmeter V to full-scale deflection. Now the short circuit across A and B is removed and the unknown resistance R_x is connected. The reading on the voltmeter V obviously will fall. It can be brought back to full-scale deflection again by moving the slider of the potential divider P farther up towards the positive end. In such an instrument it is not necessary that the voltmeter scale should be calibrated in terms of the unknown resistance—all it has to do is to indicate full-scale or some set deflection—but the position of the potential divider slider, as shown by a dial, may be so calibrated. By suitable choice of battery voltage and by varying the values of R, a wide range of unknown resistance may be easily and fairly accurately measured.

Similar principles are employed in the a.c. version shown in Fig. 16 (*b*). Here, the "battery" is replaced by power from the mains, transformed down to a voltage suitable to be applied across a continuously variable a.c. potential divider L_1 consisting essentially of a single-layer winding on a heavy circular iron core. Over this single layer moves a sliding contact arm. The resistance R of the d.c. circuit is represented by the inductance L_2 of the primary of the meter (rectifier type)

current transformer, the voltage across which is indicated by the meter M. If both A-B and C-D terminals are short-circuited, the slider of L_1 can be adjusted to make the meter M read to some set mark on its dial. Now the short-circuit is removed from A-B and the unknown inductance is connected. The reading on the meter will fall, and can be brought back to the set mark again by adjusting the slider of

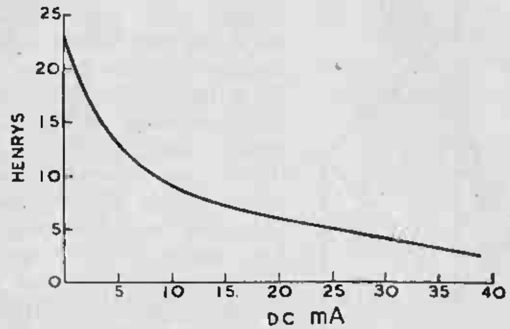


FIG. 17

A typical curve showing variations in inductance with magnetic polarization of the iron core by d.c.

L_1. The position of L_1 as indicated on the dial is calibrated in terms of the unknown inductance in henrys. By selecting taps on the primary of the meter transformer L_2 (which corresponds to varying R in the d.c. circuit) a wide range of inductance can be measured. Moreover, if the short-circuit across the terminals C-D is replaced by a battery, ammeter, and regulating resistance, a known d.c. can be passed through the unknown inductance windings, and the resultant changes in its effective reactance and thus its inductance can be ascertained by adjusting the slider of L_1 to obtain set reading on the meter with each spot value of d.c. passed. A curve of the kind shown, for example, in Fig. 17 may thus be readily obtained.

Even with this type of instrument, the effect of high resistance in the unknown inductance windings seriously affects the validity of the calibration of the slider dial. At low A.F., however, it is possible, by taking a simple d.c. measurement of the resistance of the windings, to calculate approximately its probable effect in practical apparatus. It will be noted that the d.c. also passes through the other windings of the instrument in series with the test inductance, and may affect accuracy if excessive. Once again, a warning is necessary against connecting condensers across the terminals of such instruments, in case an unlucky chance brings about resonant conditions.

All the preceding methods of measuring inductance and capacitance suffer from two limitations. They are difficult to carry out on small inductances and capacitances, and they do not reveal the power factors of condensers or the Q of coils. To deal with such measurements it is usual to make use of a.c. bridges. These are merely variations of the classic Wheatstone bridge principle with adaptations to suit the type of measurement being made. For the purpose of reference, the Wheatstone bridge principle is shown in resistive form in Fig. 18. When

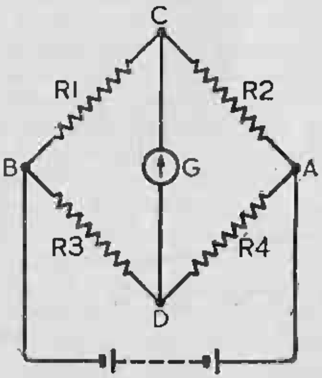


FIG. 18

The classic Wheatstone bridge circuit, still employed in laboratories and workshops for accurate measurement of resistance

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

the potential of C in relation to either A or B is the same as that of D; hence no p.d. exists between C and D, and no current flows through the galvanometer G, which registers zero. This is the condition of "balance." If it is assumed, for illustration of the practical use of this circuit, that R_1 is an unknown resistance connected for test purposes, and that $R_3 = R_4$ or $\frac{R_3}{R_4} = 1$, then balance can be obtained when $\frac{R_1}{R_2} = 1$, or $R_1 = R_2$. R_2 may consist of a standard resistance box so that when it is made equal to R_1 (as shown by the galvanometer giving zero reading), it reveals the resistance of R_1. If R_3 is made, say, 10 times as great as R_4, then R_1, in balanced conditions, is 10 times the resistance shown on the standard resistance box. Thus a very wide range of resistance can be very accurately measured.

Bridge Operation

When the bridge is energized by a.c. it becomes possible to substitute capacitances for R_1 and R_2, as shown in Fig. 19 (a). Again the basic equation holds good, $\frac{C_1}{C_2} = \frac{R_1}{R_2}$ when no p.d. exists between C and D, as is shown, in this circuit, by either

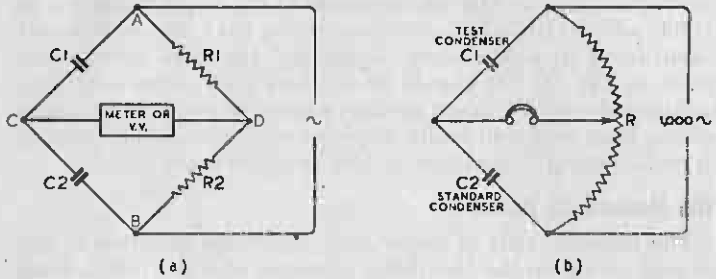


FIG. 19

When pure capacitance is substituted for resistance in the Wheatstone bridge circuit, the basic equations still hold good. In practice, C_2 is a condenser of very small or negligible power factor and accurately known capacitance

an a.c. meter or a valve voltmeter. The frequency and voltage of the a.c. energizing source do not matter apart from their influence on the choice of a balance indicator. If a bridge of this type is energized by 50 c/s mains, it is not too easy to hear in a pair of headphones (unless they have quite exceptionally good low-frequency response) when balance is obtained. It is more usual to employ some simple form of valve voltmeter, which is more sensitive than a rectifier type a.c. meter for this work. On the other hand, if a 1000 c/s oscillator is used, the note is very easily heard, and as good headphones of high resistance (inductance, really) are amongst the most sensitive forms of indicator obtainable, they are quite suitable.

Since $\frac{C_1}{C_2}$ equals merely the ratio of R_1 to R_2, a linear potential divider may be substituted for these two resistances, as in Fig. 19 (b), and balance obtained by moving the slider over R until the silent point is found. The actual resistances of the two parts of the potential divider on either side of the slider need not be known if it is a linear component—i.e. the resistance is constant per unit length—so that the physical position of the slider when at the point of balance is sufficient to indicate the ratio of C_1 to C_2; hence if C_2 is a standard condenser of known capacitance, the actual value of C_1 can be ascertained from the position of the slider and the dial thus calibrated directly in terms of the test capacitance. These forms of capacitance measuring bridge are quite satisfactory at audio frequencies and for the measurement of capacitances as low as 0·0005 μF. With a little common sense they are quite easily constructed in a workshop, and some practical circuits are given in Fig. 20. It should be obvious that elementary precautions should be taken against excessive stray capacitance arising from long and badly disposed wiring and its proximity to metal casing. Insulation is also of importance.

The Resistance Factor

The measurement of inductance by bridge methods is not so easy owing to the inevitable presence of a d.c. conducting path through the wire and its resistance. This will be dealt with later, but this factor of resistance perhaps existing in addition to reactance crops up also in the measurement of capacitance, especially with large condensers and with small condensers used in R.F. circuits. It will be noted that in the practical circuits given in Fig. 20 variable resistance is included in series with the largest values of standard condenser. This resistance is intended to simulate the effect, in the arm containing the standard low-loss condenser, of resistance or losses connected with the test condenser. It is not included with the smaller standard condensers, because "power factor" is very seldom of great consequence when measuring these comparatively large values of capacitance that are to be used in A.F. circuits—unless the condenser being measured has a defect, such as a leak, when this is tested for more directly by plain d.c. resistance measurement methods.

In some commercial bridge instruments for measuring capacitance, the variable resistance is actually calibrated, not in terms of the resistance, but in terms of the power factor of the condenser under test. If the test condenser has a "bad power factor," the effect is, in a simple bridge of the type in Fig. 19 (*a*) and (*b*), to make the balance point indeterminate, so that complete silence is not obtained in the headphones or zero reading in the indicating meter. It is as if the test

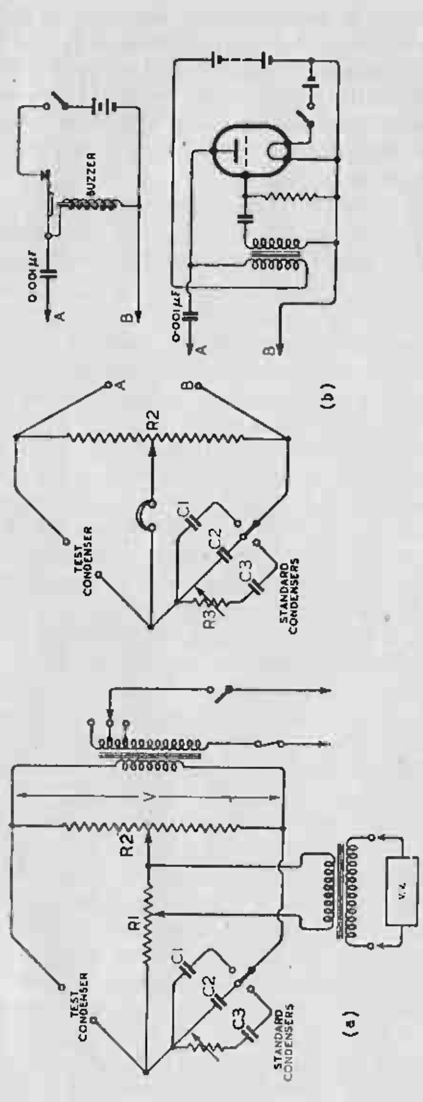


Fig. 20

(*a*) Practical form of bridge energized by a.c. 50 c/s mains from a step-down transformer. Typical approximate values are V = 20 volts, R_1 (sensitivity control) and R_2 = 10,000 ohms, C_1 = 0·001 μF, C_2 = 0·1 μF, C_3 = 1 μF, which enables from 0·005 μF to 10 μF to be measured with usable accuracy. (*b*) Simple bridge circuit which may be energized either by a high-note buzzer or a valve oscillator using an intervalve transformer. R_2 and C_1, C_2, C_3 are of the same values as in (*a*). R_3 may be about 1000 ohms, and adjustment of it may be found necessary to obtain sharp balance when testing large condensers of poor power factor condenser had a resistance in series with it, which would alter the phase relationship between the current through it and the voltage across it from the exact 90° that would occur if it were a pure capacitance. (A similar effect would occur if there were capacitance across one of the resistances in the other arm of the bridge; hence the importance of reducing the value of stray parallel capacitance to a minimum.)

The subjects both of inductance measurement and of small capacitance measurement, especially at R.F., and by methods other than bridge, will be dealt with in the next chapter.

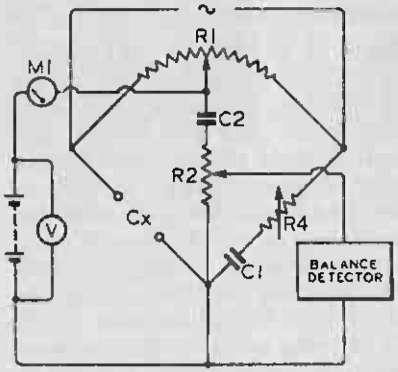
CHAPTER VI

ELECTROLYTIC CONDENSER TESTING AND INDUCTANCE AND CAPACITANCE AT R.F.

In Chapter V the measurement of inductance and capacitance by testers and bridges utilizing mains and A.F. power supplies was dealt with. It was found possible to measure inductance between about 0·5 henry and 100 henrys, and capacitance between 0·0005 μF and 50 μF, by such means. Before going on to the matter of measurements at R.F., of very small values of inductance and capacitance, the measurement of electrolytic condensers may as well be cleared up.

In an electrolytic condenser the dielectric consists of a very thin layer of oxide on an aluminium foil electrode, created by "forming," i.e. passing current from the aluminium to the solution (contact with the latter being made by another piece of aluminium or the outer case). Since the oxide layer can be made extremely thin, an electrolytic condenser can be made to have a very high capacitance for its size, compared with other forms of condenser. However, the oxide layer is not stable; it is rapidly broken down if the direction of the current through the condenser is reversed. If it is to handle alternating voltage, it must have a steady d.c. potential impressed on it so that at no time do the negative peaks of the alternating voltage exceed this d.c. potential; in fact, the alternating voltage must only be a comparatively small fraction of the d.c. if overheating and breakdown are to be avoided. Finally, the power factor is always very much higher than paper or mica dielectric condensers. This does not matter greatly at mains and audio frequencies, which is why electrolytic condensers are so often employed as smoothing condensers.

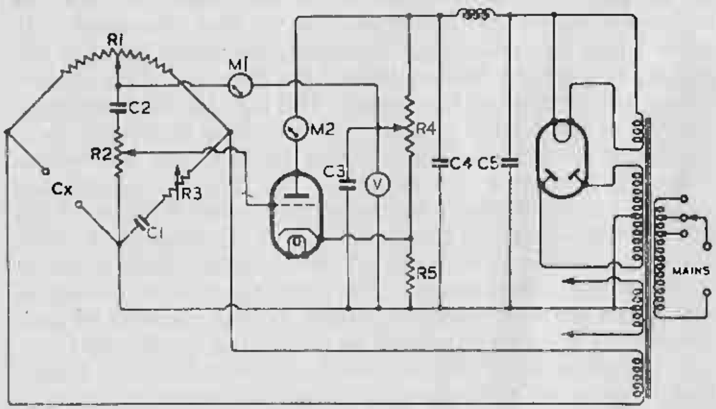
It is evident that they cannot be measured on the testers that have so far been described, since no provision has been made for the introduction of a steady d.c. polarizing voltage. The form of bridge shown in Fig. 21 (*a*), however, enables a polarizing voltage to be applied without affecting the action of the bridge, and by the inclusion of a milliammeter M_1, as shown, the leakage current can also be measured. The goodness of an electrolytic condenser is seldom given in terms of the "power factor," because this is always very high, but usually in terms of mA per μF "leakage." Thus a 10 μF condenser may, with its working voltage impressed across its terminals, have a leakage current of 0·1 mA, and the goodness will then be expressed as "0·01 mA/μF" at a given voltage.



(a)

FIG. 21

(*a*) Circuit for testing the capacitance and leakage of electrolytic condensers, using batteries for the polarizing voltage. 50 c/s from a mains transformer low-voltage secondary is a convenient and suitable means of energizing the bridge, and a sensitive a.c. milliammeter or valve voltmeter should be used to detect balance. (*b*) A mains version. R_3 is the sensitivity control, M_1 the leakage meter, and M_2 the balance detector in the anode circuit of the simple valve voltmeter. Smoothing must be extremely good



(b)

Since many electrolytic condensers are rated at several hundred volts (about 600 is roughly the maximum) it is advisable to employ mains supply power to energize the bridge and provide the polarizing voltage, and for those who have a number of electrolytics to test it is worth while making up a

mains-driven instrument for the purpose on the lines shown in Fig. 21 (*b*).

The balancing resistance R_3 shown in the standard condenser arm is mainly for the purpose of securing a satisfactory zero indication of balance in view of the high power factor of the electrolytic being tested. Since the bridge is energized by a winding on the mains transformer at 50 c/s, a simple valve

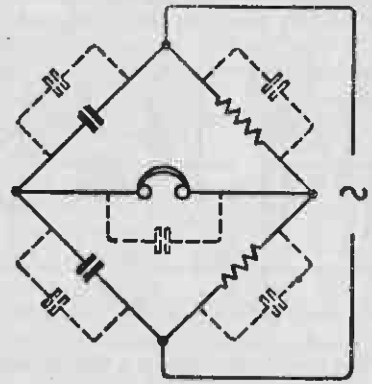


FIG. 22

The existence of stray and unknown capacitances in a simple bridge makes it unreliable for measurements of low values without special precautions

voltmeter is usually better as a balance indicator than headphones.

R.F. Measurements

The chief difficulties in measuring small values of inductance and capacitance by means of bridges arise from the very high or low reactances offered by them, and from the inevitable presence of stray reactances due to the various components of the bridge. For instance, a 10 $\mu\mu$F condenser offers a reactance at even 1000 c/s of 16 million ohms. In the first place, a bridge with reactance of such an order in each arm would require an inordinately high energizing voltage to enable balance to be accurately obtained. In the second place, if the reactance were to be reduced by increasing the frequency of the energizing voltage, there would still be the stray capacitances to be reckoned with, as shown in Fig. 22.

Bridges that are energized by R.F. at, say, 1 Mc/s are, therefore, constructed with elaborate and careful screening that aim at making all the strays symmetrical. This makes the instrument costly, and, unless repeatedly checked, none too reliable. Balance is indicated by means of some form of valve voltmeter or oscillating detector, which adds further to the complication.

However, a R.F. bridge in a special form may be used to match two variable condensers on one shaft. The test condensers themselves are made to form one side of the bridge,

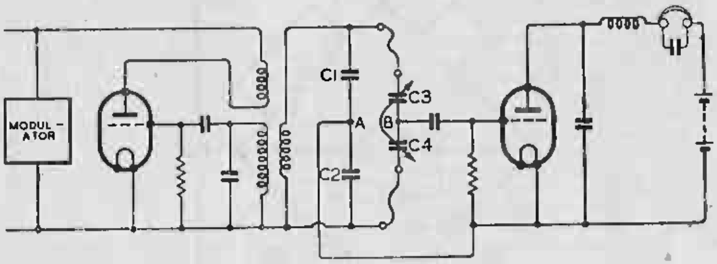


FIG. 23

A bridge for matching ganged condensers (C_3C_4). Very careful matching also of the stray capacities is essential

as shown in Fig. 23. Two equal standard condensers, or resistances, form the other side of the bridge, and as the gang condenser under test is rotated, any inequality in the capacitances of the two halves is revealed by the development of a p.d. across A and B—i.e. balance is lost. It may be restored, of course, by adjustment of, say, split end-vanes in one or the other of the two ganged condensers.

The almost standard method of measuring very small capacitance is the substitution method, using a carefully calibrated variable standard condenser—a glorified "straight-line-capacity law" tuning condenser similar to a type that used to be employed in receivers at the very beginning of broadcasting before "log-law" and "straight-line-frequency" types came much into popular use. Such variable standard condensers are expensive and, unless extremely well made, apt to suffer from loss of accuracy due to temperature and other effects, but they are amongst the most useful pieces of apparatus in the laboratory. Their capacitance varies proportionally to the amount of meshing of stator and

rotor vanes, and the scale is calibrated to read directly in $\mu\mu$F.

Substitution Methods

One way of using such a condenser for the measurement of a small value of capacitance is shown in Fig. 24. The variable standard condenser C_1 is made to act as the tuning capacitance in a simple valve oscillator. The radiation from this oscillator is received on a simple form of oscillating receiver—the plain reaction type is suitable—which is tuned to zero beat with the testing oscillator. Under these conditions, the capacitance of the standard variable condenser is noted from its dial, and then the small condenser C_2 to be tested is connected with very short leads across the standard. This will alter the frequency of the testing oscillator, and the oscillating receiver will then produce a beat-note in the headphones—unless the change in tuning capacitance brought about by the addition of the unknown condenser across the standard is so great that the beat-note is too high for audibility. In any case, the standard variable condenser is now altered in capacitance until zero beat is again obtained in the oscillating receiver phones, and the dial setting (or value of capacitance) required to do this noted. Then, plainly, the value of the unknown capacitance is equal to the difference between the first setting of the standard variable and the second setting. With care, this

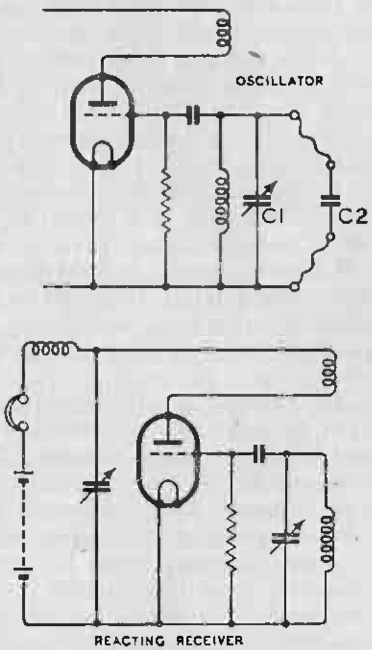


FIG. 24

Set-up for measuring a small capacitance, C_2, by noting its effect on the tuning of the oscillator and consequent necessary alteration in the setting of the calibrated variable standard C_1

can be a very accurate method of measuring small capacitance. The principle can be used, of course, to match the sections of a ganged condenser; first, the testing oscillator is tuned by one section of the gang and zero beat obtained in the checking oscillating receiver, and then the lead is changed over to another section of the gang (without, of course, altering the setting of the gang common shaft), and if zero beat is obtained, the sections involved are of equal capacitance.

Before going on to the matter of measuring small coil inductance, mention must be made of two factors met with in R.F. work that have much more importance than in ordinary A.F. measurements. In connection with condensers it is the "power factor" and in connection with coils it is Q. It is important to realize that both these factors come into play in estimating the performance of a circuit at high frequencies as a result of a peculiar thing that is summed up in the words "H.F. resistance." A slight digression may here be helpful to what follows later. It should be understood that the essential effect of resistance, as differing from reactance, lies from a practical point of view in the loss of energy that it brings about.

For example, a condenser may have a high "power factor" even though the insulation between the plates is excellent and the capacitance normal. This may occur because the dielectric material is unsuitable. When a condenser is repeatedly charged and discharged and then charged up with reversed polarity and again discharged and so on—which is what happens when a condenser "passes a.c."—the dielectric layer is subjected to cyclic strains in its atomic structure. The effect has an analogy in the compression and expansion of a piece of rubber. Suppose, however, that putty is used instead of rubber. Unlike rubber, when the putty is compressed it stays compressed or flattened and does not give back the energy required to flatten it. The energy devoted to the flattening and expansion manifests itself eventually as heat. An analogous sort of thing occurs in the dielectric of a poorly made condenser, and the final manifestation is again heat, which may eventually break down the insulation between the plates. It is for this reason that electrolytic condensers, although they can withstand a considerable steady d.c. potential between their electrodes, break down if too high and rapid an alternating voltage is superimposed on this d.c. potential.

Thus it may be plain how *losses* can occur in a condenser without any obvious "Ohm's Law" resistance being present. Moreover, the higher the frequency involved the greater the losses.

It is convenient to express these losses in terms of their electrical effects, which are the same as those of actual "Ohm's Law" resistance. One of these effects is a departure from the ideal 90-degree phase relationship between the current through and the voltage across a condenser passing a.c.—hence the apparently odd term "Cos θ" in which the "power factor" of a condenser is very often stated. This expresses the ratio between the true power and the apparent power in the circuit in terms of the phase angle θ between voltage and current. In a perfect condenser θ is 90 degrees and "Cos θ" is zero.

Defining "Q"

Losses can also occur in a coil, independently of the "Ohm's Law" resistance of the windings of the coil, which may be, in the case of a short-wave tuning coil, very small indeed. For example, a coil inside a screening can inevitably causes more loss of electrical energy than a coil without a can and mounted well away from conductors such as metal chassis and screening plates, although the "Ohm's Law" resistance as measured with d.c. is no greater. This is because the alternating current through the coil sets up a moving magnetic field that induces current in the screening can, and this current in the screening can material, which always offers some resistance, means a transference of electrical energy from the coil to the can from which it cannot be recovered.

If the foregoing very sketchy outline of "H.F. resistance" has been understood, the meaning of Q may be clear from what follows. Q is a term that has come to be accepted as a "figure of merit" for coils and tuned circuits. It is the ratio of the purely reactive impedance of a circuit to the resistance—resistance, in this case, and especially at R.F., meaning the a.c. and not the "Ohm's Law" resistance. For a coil, therefore, $Q = \frac{\omega L.}{R}$ For a tuned circuit containing both inductance and capacitance, it does not matter whether the resistance exists (or the losses occur) in the inductance or the capacitance so far as the behaviour of the circuit as a whole is concerned, and it is therefore possible, in theory, to regard Q as still dependent on the goodness of the coil. This is actually nearly always the case, since, except at the ultra-high frequencies, good air-spaced condensers have very low losses as compared with even the best of coils.

The measurements that follow, of the dynamic resistance of a tuned circuit, the inductance and the Q of a coil, the comparative goodness of a condenser, and the self-capacitance of a coil, are carried out by means of apparatus that is not quite

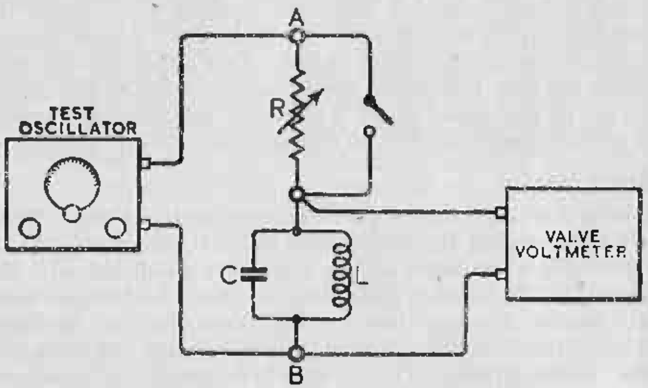


FIG. 25

Set-up for measuring the dynamic impedance of a parallel tuned circuit, LC. The variable resistance R must be of very low self-capacity and residual inductance. Some composition types are suitable even at intermediate frequencies

so simple in either construction or use as that which has been discussed already. The accurately calibrated variable condenser, of the order of 0·0005 μF maximum, is practically essential, and there are also required an accurately calibrated test oscillator and a good valve voltmeter. These three items are actually included in a form of composite instrument known as a "Q-Meter" that is available in several commercial forms. For present purposes, however, it may be best to regard them as separate instruments.

The dynamic impedance of a tuned circuit may be measured in several ways, of which only two will be dealt with—the "equivalent resistance" and the "negative resistance" methods. The principle of the first is simply that an actual ohmic resistance is placed in series with the tuned circuit and adjusted

until either the make-up current through, or the voltage developed across, the tuned circuit is halved. The set-up is illustrated in Fig. 25. Taking the voltage method first, the voltage developed across the tuned circuit LC by the test oscillator is measured by the valve voltmeter with the series resistance R short-circuited. Then the resistance is brought into circuit by the removal of the short, and adjusted until exactly half the first voltage is shown on the valve voltmeter. Then, plainly,

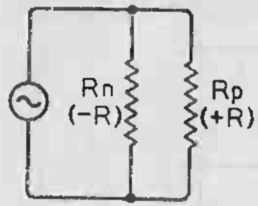


FIG. 26

Showing the principle of adding positive (R_p) and negative (R_n) resistances in parallel

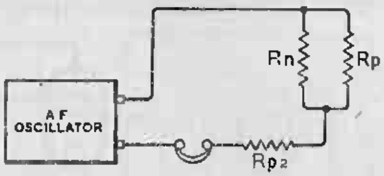


FIG. 27

General form of circuit for measuring the value of an unknown negative resistance R_n. The positive resistances, R_p, are familiar as "ordinary" physical elements. When R_{p1} is equal to R_n, the parallel combination of R_n R_{p1} offers infinite resistance, hence no sound is heard in the phones. R_{p2} is a safeguarding resistance. R_n may be a valve circuit such as a dynatron

the value of the ohmic resistance, which can be ascertained by the usual d.c. methods separately, is equal to the impedance offered by the tuned circuit. The usual precautions against stray capacitance due to wiring and to the resistance R itself have to be guarded against.

Negative Resistance Methods

A "negative resistance" is a device that supplies energy to a circuit—the opposite of a positive resistance, that is. It can be represented in a circuit by means of the usual symbol for resistance, and suitably marked to distinguish it from positive resistance. In Fig. 26 are shown a negative and a positive resistance in parallel, while a generator applies a p.d. across them. The effective resistance of this parallel combination is given by $R = \dfrac{R_p \times R_n}{R_p + R_n}$. Now, R_n is a minus quantity; hence, if $R_p = R_n$, $R_p + R_n = 0$ and then $R = \dfrac{R_p \times R_n}{0}$, and becomes infinite. No current will then flow through the parallel combination of R_p and R_n.

A circuit can be made up, therefore, to measure the ohmic value of a negative resistance, of the general form shown in Fig. 27. R_{p1} is adjusted until no sound is heard in the phones, which means that $R_{p1} = R_n$. R_{p1} may then be measured by ordinary means, using d.c., unless, being a calibrated variable resistance, its value is already known from the dial reading.

There are several devices that provide negative resistance,

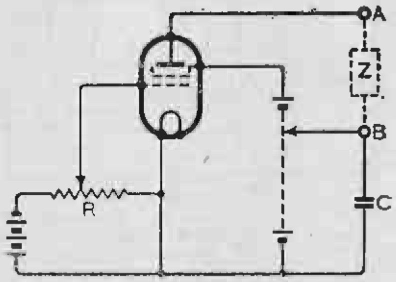


FIG. 28

A screened grid valve used as a dynatron or as a negative resistance across *A* and *B*. The value of the negative resistance so produced may be controlled by a negative bias on the control grid. *Z* may be a tuned circuit. Note that the anode is at a lower h.t. voltage than the screening grid. This is a fairly critical adjustment. *C* is a by-pass condenser of large value

of which the valve used as an oscillator is one, and the form of oscillator that is chiefly used in laboratory measurements is the "dynatron," on account of the ease with which tuned circuits to be tested may be connected (no "reaction coil" is needed) and the purity of the waveform it generates.

This is simply a screened grid valve, connected as shown in Fig. 28, and the tuned circuit to be tested is connected to the test terminals *A* and *B*. For the operation of this device, reference should be made to Fig. 29. The process—with variations due to any particular form in which the principle is employed in a practical instrument—is in general terms as follows: First, the negative resistance of the valve is adjusted by means of the grid bias until the tuned circuit is *just* set into oscillation. This is detected by a simple form of autodyne or oscillating receiver, in the phones of which a beat-note can be obtained when the tuned circuit under test is oscillating. Now

the tuned circuit is disconnected and in its place is connected the circuit for measuring the screened grid valve negative resistance, which—since this was made just sufficient to maintain oscillation in the tuned circuit—will be the same as that of the tuned circuit.

Measuring Inductance

The inductance of a coil of the order of millihenrys can be measured by a bridge circuit, if a calibrated variable induc-

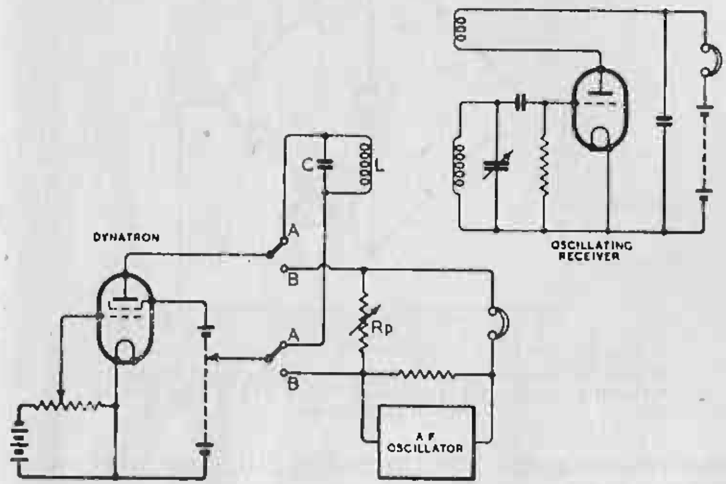


FIG. 29

Skeleton circuit showing general principles for measuring the dynamic impedance at resonance of a tuned circuit *LC*

tance is employed, as shown in Fig. 30, and the bridge is energized by a fairly high A.F., say of the order of 1000 c/s, from an oscillator. An inductometer provides the required standard variable inductance, and consists of a glorified version of what used to be called a "variometer"; two inductances are connected in series and arranged so that one is fixed and the other is rotatable so that its field may either assist or oppose that of the fixed inductance; the overall inductance of the two coils in series can thus be varied from zero to their sum with the mutual inductance aiding. But this bridge method is not satisfactory with inductances below a few millihenrys, and recourse must then be had to first principle methods. A calibrated test oscillator, to energize the tuned circuit formed by the coil under test and a calibrated variable condenser, and a valve voltmeter to indicate resonance, are employed. The process is to set the condenser at some definite capacitance, and to connect the coil under test across it. Then the test oscillator is coupled to this tuned circuit in such a way that it

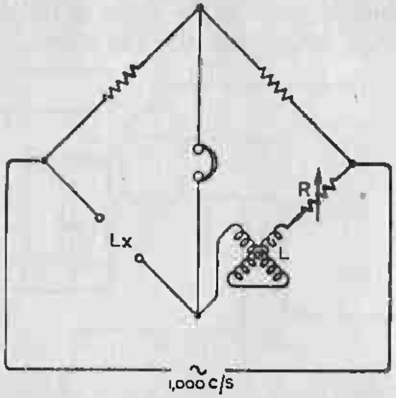


FIG. 30

Bridge for measuring an inductance L_x. The effect of the d.c. resistance of this coil is balanced by R, and L is a variable standard inductance

does not also couple with the pick-up coil of the valve voltmeter.

The test oscillator is then tuned to resonance with the tuned circuit which is indicated by maximum reading in the valve voltmeter. Knowing the frequency from the calibration of the test oscillator, and the capacitance of the condenser, the inductance of the coil can be worked out from the formula $f = \frac{1}{2\pi\sqrt{LC}}$ from which $L = \frac{1}{4\pi^2 f^2 C}$ where L is in μH, C in μF, and f in Mc/s.

The Q of a coil is measured by making use of the fact that the voltage developed across a coil in a series circuit with a condenser is Q times V, where V is the applied voltage. Reference to Fig. 31 (*a*) shows the principle. First, the valve voltmeter measures the voltage applied across the series tuned

circuit *at resonance*. At resonance, the only impedance offered by such a series tuned circuit is due to the resistance—which may be actual ohmic resistance due, say, to the resistance of the coil wire, or to some form of loss, as was explained earlier, such as the presence of a screening can. Let this applied voltage be V_1, and the voltage across the coil L be V_2. When measured by the valve voltmeter, V_2

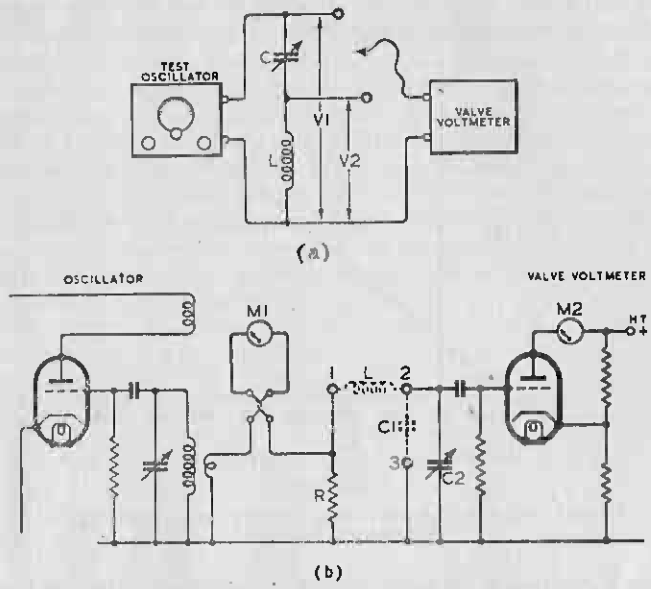


FIG. 31

(*a*) Circuit showing principle of coil Q measurement. (*b*) Basic circuit of a Q or Circuit Magnification Meter. M_1 is a thermo-couple R.F. meter used to develop a standard output from the oscillator across R the coupling resistance. The test inductance L is normally in series with the standard variable condenser C_2, but across 2 and 3 a condenser under test may be connected

will be found much greater than V_1—in fact, it will be $Q \times V_1$. Because $V_1 = IR$ and $V_2 = I\omega L$, where I is the current through the tuned circuit. Hence

$$\frac{V_2}{V_1} = \frac{I\omega L}{IR} = \frac{\omega L}{R} = Q$$

If, now, V_1 is made equal to 1 volt, the voltage V_2 measured on the valve voltmeter is a direct indication of the Q of the coil. Since Q may be as high as 100, necessitating the use of a valve voltmeter capable of measuring 100 volts, and also in order that the impedance of the output of the test oscillator, which acts as a generator, may be so low that it can be neglected, a Q-meter usually consists of an oscillator developing a low voltage of, say, 0·1 volt across a very low resistance which is in series with the tuned circuit, as shown in Fig. 31 (*b*). Then a Q of 100 would be shown as 10 volts on the valve voltmeter scale, and so proportionately with other values of Q. If the

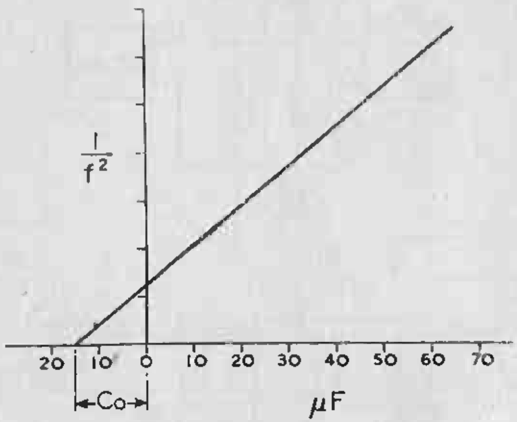


FIG. 32
Graph of frequency against tuning capacity, for ascertaining the self-capacity of a coil

tuning condenser happens to be a calibrated standard, of extremely good electrical construction—i.e. air-spaced and losses so low that they may be ignored in all but very high frequency circuits—the goodness of small test condensers with, say, solid dielectric may be tested for their effect on the Q of the tuned circuit by placing them in parallel with the standard condenser and observing the difference they make to the Q indicated on the valve voltmeter.

The self-capacitance of a coil is again ascertained by using a series of known frequencies from the test oscillator and the standard variable condenser. The procedure is to set up the apparatus as if for measuring the inductance of the coil, as has been described earlier. Then a series of observations are

taken of the resonant frequencies obtained with known small increases of tuning capacitance. From these two sets of figures —frequency and tuning capacitance—a graph is plotted, as shown in Fig. 32, and the line made by joining the points is extended beyond the vertical ordinate. Where this extended line cuts the horizontal or capacitance ordinate gives the residual capacitance due to the self-capacitance of the coil itself.

The foregoing presents a brief outline of a branch of measurement that has already a very wide choice of methods, and that is presenting fresh problems as the study of higher frequency phenomena expands. It will be seen that most of the measurements involve methods and precision instruments of the laboratory, rather than the ordinary workshop, type, and of deductions from first principles. There is considerable scope in this field for the ingenuity of the "practical man" in devising simple and easily operated devices and circuits for use with the ever higher frequencies that are becoming involved in modern wireless technique.

CHAPTER VII

BEAT FREQUENCY OSCILLATORS

In testing the performance of audio amplifiers and receivers it is often necessary to make use of audio frequency alternating voltage of which the frequency, the magnitude, and the waveform are precisely known. Moreover, it is very often necessary that the first two factors at least should be easily and accurately altered over wide ranges.

For instance, the response of an A.F. amplifier may be checked by applying a test input voltage and recording the resultant output on an output meter. In such a case the input voltage may need to be less than a volt in magnitude, but that magnitude must be known; the *frequency* of this voltage, however, must be variable from (for a first-class amplifier) about 20 c/s to 12,000 c/s, and in most cases it is very important that it should be of pure sine waveform. A known voltage of this nature, but of much greater amplitude, may be needed to modulate some test R.F. carrier signal when measuring the performance of a receiver. There is, again, the testing of a loudspeaker for its efficiency and fidelity in converting electrical into acoustical energy, and for this quite considerable power, as well as accurately known frequency and waveform, may be needed. A little thought will show that a piece of apparatus capable of producing alternating potentials of such widely variable qualities with ease and certainty has to be a fairly complex device.

Sum and Difference Frequencies

Indeed, a good beat frequency oscillator, which is the most commonly used instrument for the purpose, is neither easy nor cheap to construct. Its principles, however, are simple enough. It is well known that if two sine-wave alternating voltages are developed simultaneously across some circuit that will record their resultant effect, that effect will contain not only the two original frequencies but also two more, one equal to the sum of the two originating frequencies and the other equal to their difference. Thus if $f1$ and $f2$ are the two originating frequencies, the mixing of them will give rise to $f3 = f1 + f2$ and $f4 = f1 - f2$.

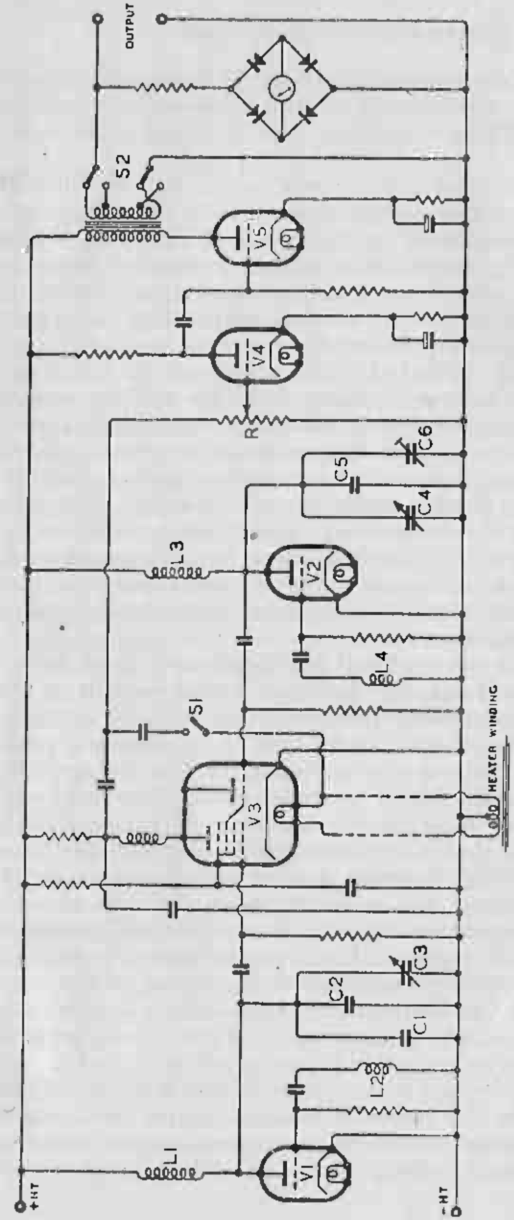


Fig. 33

Basic circuit of a typical beat-frequency oscillator. V_1 and V_2 are tuned anode oscillators. C_3 = 0·00005 μF (variable), and is used as the "Set Zero" control in adjustments preliminary to use. The output frequency is altered by C_4 = 0·0005 μF. C_6 = 0·00005 μF and is a pre-set trimmer for scale calibration. $C_2 = C_5$ = 0·001 μF; these form the major part of the capacities tuning the circuits

To take some figures: if $f1 = 100$ kc/s and $f2 = 115$ kc/s, then the resultant of mixing them will contain also $f3 = 100 + 115$ kc/s $= 215$ kc/s, and $f4 = 115 - 100$ kc/s $=$ 15 kc/s.

A glance at the circuit diagram shown in Fig. 33 will reveal two triode tuned-anode oscillators, V_1 and V_2, the frequencies of which are determined by the values of L_1, C_1, C_2, C_3, and L_3, C_4, C_5, C_6, respectively. The outputs of these two oscillators are mixed in a conventional triode-hexode, so that in the anode circuit of that valve there are present the four frequencies, two from the two oscillators, their sum, and their difference. The frequency of the oscillators themselves is approximately 100 kc/s, and the sum frequency is, therefore, approximately 200 kc/s. These three are of comparatively high frequency and are readily filtered out by chokes and by-pass condensers, leaving only the difference frequency, which is within the audible range. This difference frequency is now handled exactly like an ordinary A.F. input to the grid of V_4, which acts as a buffer amplifier, and the signal passes to the output valve V_5, and is conveyed to the output terminals through a massive and carefully designed iron-cored transformer.

This is a fairly typical and very commonly used form of circuit for a beat frequency oscillator. Before considering other forms of A.F. generator, it will be worth while to note a few points about the circuit. In the first place, although it is not shown in the diagram, the screening of the two oscillators from each other has to be very complete. They are usually positioned as far away from each other as is practicable in the instrument and the coils at least are encased in cans. This is to avoid "pulling" between the two oscillators when they are on very nearly the same frequency. Those who have handled a "reaction" receiver close to a powerful transmitter will have noticed, perhaps, that when the receiver is oscillating the resulting whistle caused by the heterodyning of the incoming signal with the oscillation of the reacting valve cannot be brought progressively quite to zero. When it has dropped to a few score cycles per second it stops abruptly and a "dead spot" seems to extend over a degree or two of the tuning dial. This is because the powerful incoming signal is forcing the oscillatory currents in the receiver tuned circuits into resonance with itself. Plainly, when the two frequencies of the

incoming signal and of the oscillatory currents in the receiver are exactly the same, there is no difference frequency—that is, no audible sound.

Hexode Mixer

The same sort of thing may happen if the two oscillators in the B.F.O. pull into resonance, through stray coupling, at a common frequency; they produce in the anode of the mixer valve V_3 no difference frequency, or a zero beat frequency. This would mean in practice that the B.F.O., instead of being able to produce alternating output voltage at frequencies covering even the lowest audio extreme (about, perhaps, 30 c/s), would give no output at all below about 100 c/s or so—which would render the B.F.O. useless for quite a number of the more interesting performance tests.

Hence, the two oscillators are most carefully isolated from each other. The use of a triode-hexode mixer aids this—it is a type of valve noted for the small amount of coupling between the triode portion and the control grid circuit even in normal use in superheterodyne reception—and the very fact that the triode portion of this valve is not used save for the triode grid, but the two frequencies are generated by separate oscillators, enables them to be brought to within very small differences from each other before they tend to pull into resonance.

There is also a reason for picking on neither a very high nor a very low frequency for the two oscillators. Considering first what would happen if V_1, say, were fixed at 40 kc/s and V_2 had to be tunable to produce difference frequencies covering the whole A.F. range, it is evident that to do so V_2 would have to be tunable from 40 kc/s to 25 kc/s. It would itself run nearly into the A.F. range, and an enormous tuning condenser would be required to cover this range of frequency variation. The oscillator frequencies are therefore made high enough to be well outside the A.F. range and easily tunable over 15 kc/s, using the almost standard 0·0005 μF type of tuning condenser in conjunction with fixed condensers and trimmers.

Optimum Oscillator Frequency

Now, if we consider what would happen if, instead of about 100 kc/s, we were to use frequencies of the order of megacycles per second, it is found that the capacity changes to cover the

15 kc/s range of difference are so slight that they can be brought about by small alterations of stray capacities due to wiring, changes of valves, warping of the case, etc. So that with two high-frequency oscillators, tuning would certainly be easy, but the order of reliability and constancy would be remarkably low; moreover, owing to the small percentage of difference between two such high frequencies, pulling between their circuits would become much more difficult to avoid. Hence, 100 kc/s seems to be a good working compromise.

In the diagram, V_1 is the fixed oscillator, the trimmer C_3 shown being used only for zero adjustments. V_2 is tuned over a range from 100 kc/s to 115 kc/s by the 0·0005 μF variable condenser C_5, across which there is also a fixed condenser C_5, which corresponds to C_1 in V_1, and it, too, has a trimmer that may be adjusted to compensate for changes in strays. The difference frequency is determined by the setting of the condenser C_4, the dial of which is calibrated in cycles per second.

Although the two voltages are combined, in this case, in a triode-hexode, another triode is often used. With careful design, one of the oscillators itself may function also as a mixer.

In the anode of the mixer, by-pass condensers and filters remove the supersonic components present, leaving the difference frequency, which is then fed to the grid of a buffer-amplifier valve stage, and treated as an ordinary A.F. signal. There may be more than one of these stages.

A point worth noting is that connected to the grid of the buffer-amplifier there is not only the output control potential divider, but also a switch S_1 that connects this grid at will to one side of the valve heaters. This is done in order that, when setting up the instrument for use, a quick check of accuracy of calibration may be made against the 50 c/s frequency of the a.c. mains supply that operates the instrument. Normally, "mains hum" is carefully avoided in amplifiers, one of the devices for this purpose being to centre-tap the heater supply windings as shown in Fig. 33 to chassis. However, by connecting the grid of V_4 through a condenser to one side of these heater windings, a volt or two at 50 c/s is applied.

Now, if the difference beat frequency supplied from the anode of V_3 is itself at 50 c/s, it will be exactly in step with this mains frequency and the pointer of the output meter of the instrument will merely indicate a steady output. But if the generated

difference frequency is not exactly 50 c/s, the pointer will slowly rise and fall, and the "Set Zero" condenser (C_3) must be adjusted until this waving motion ceases and the generated and mains voltages applied to the grid of V_4 are exactly in step. Admittedly this check can only be applied at the 50 c/s mark on the calibrated dial, or, if harmonics are sufficiently strong in the mains supply, at 100 and 150 c/s; but if the rest of the instrument is in good order, it can usually be assumed that the calibration is correct over the rest of the dial.

It will be noted that the only iron in the circuit is in the output transformer, which is usually of massive construction to avoid an approach to saturation, even on high outputs, with consequent distortion and the creation of harmonic frequencies in the output. Provision is shown for only two output impedances obtained by altering the ratio of secondary to primary in the output transformer; but, of course, the design can be adapted to the nature of the work to be done. A rectifier type of a.c. voltmeter is usually quite accurate enough to indicate the voltage developed across the output terminals.

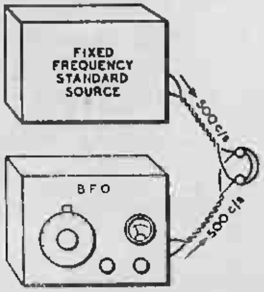


FIG. 34

A method of calibration by audible beats in which a fixed frequency standard and a B.F.O. have their outputs applied simultaneously to a pair of headphones or a loudspeaker

The calibration of a B.F.O. and of kindred sources of A.F. may be carried out either by means of comparison with standard fixed generators, including tuning forks, or by visual means with a cathode-ray oscillograph. Considering the first-mentioned, all that is necessary is to combine the standard frequency output acoustically with that of the B.F.O. under test. For instance, at frequencies over 100 c/s (above which the response of headphones is usually sufficient) the arrangement of Fig. 34 will be quite satisfactory. The output of the standard source and that of the B.F.O. will combine as a complex sound output from the phones. The dial of the B.F.O. may now be turned until the B.F.O. output frequency approaches the standard frequency, say, 500 c/s. As coincidence is made closer, the combination of the B.F.O. and the standard outputs will give rise to a beat note or flutter, falling in frequency until finally only a pure 500 c/s tone is heard. In this way, accuracy of the order of a fraction of a cycle per second can be obtained with ease, even at high frequencies of the order of 10 or 12 kc/s. Even with only one standard frequency, the B.F.O. dial may be calibrated by the use of octaves and a good musical ear, with an accuracy quite sufficient for most purposes. If this method of calibration is adopted one should arrange to be at home when the piano-tuner calls!

Good public mains supplies preserve a considerable accuracy

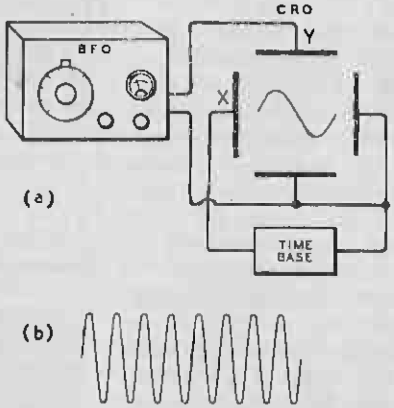


FIG. 35

(a) Shows the picture obtained when the frequency of the time base and the B.F.O. output are the same. (b) Shows the picture obtained when the frequency of the B.F.O. output is increased to eight times that of the time base. There are eight complete sine waves

of frequency, even if their voltage varies somewhat, and therefore form a very handy standard 50 c/s frequency. With this source at 50 c/s, Lissajous figures obtained on the screen of a cathode-ray oscillograph (Fig. 35) provide a very accurate and easy method of calibrating the dial of a B.F.O. over the whole A.F. range. The method may be briefly outlined as follows: First, the saw-toothed time base for the X sweep is set up at exactly 50 c/s, by applying mains voltage (stepped down by a transformer) to the Y plate and adjusting the time base until a *single* sine-wave trace is obtained. Now the mains voltage is removed from the Y plate and the output of the B.F.O. is applied in its place. The synchronizing control

should *not* be used to steady the picture; that must be done by exact control of the time base frequency and B.F.O. output.

If the B.F.O. output is now made 25 c/s, an elliptical figure will be obtained on C.R.O. screen. If desired, the time base can now be adjusted to run more slowly until the sine wave is regained, showing the time base to be set at 25 c/s. With this time base, the application of B.F.O. outputs at multiples of 25 c/s can be readily obtained by increasing the B.F.O. frequency and noting the necessarily increasing number of

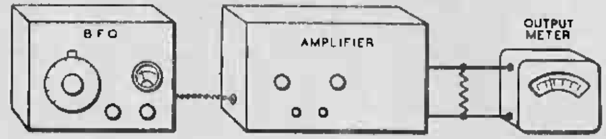


FIG. 36

An arrangement for taking a frequency response curve of an A.F. amplifier. R is a resistance equal to the normal load offered by the speaker

sine waves in the picture, up to some 200 or 250 c/s. Assuming that the last figure has been obtained with 250 c/s applied to the Y plate from the B.F.O.—10 complete sine waves on the screen—the time base is then adjusted, leaving the B.F.O. dial untouched, to increase its speed until once more only one sine wave is seen as the picture. The time base is now running at 250 c/s. Multiples of this can be obtained now, as before, by increasing the frequency of the B.F.O. output so that in succession 2, 3, 4, 5, and up to 10 complete sine waves are seen. This takes the B.F.O. up to 2500 c/s. The time base can now be again adjusted to regain a single wave, showing that it is running at 2500 c/s, and with that as a base the B.F.O. output can be taken in frequency right up to the limit of the A.F. band.

Having obtained a calibration point at, say, 1000 c/s on the B.F.O. dial, it can be used on similar principles to find a very large number of intermediate points. With care, this can be made an extremely accurate method of calibrating a frequency source up to radio frequencies, the limit being determined by the maximum speed of the time base and the characteristics of the cathode-ray tube.

The uses of a good B.F.O. are so many and varied that it is hard to select a few typical examples. There is, of course, the most obvious use, in checking over an A.F. amplifier for "frequency response." The arrangement is shown in Fig. 36, the B.F.O. output, maintained at an unvarying level, being applied to the input terminals and the frequency altered to cover the whole A.F. range desired. The variations in the output of the amplifier as recorded on the output meter are noted and subsequently plotted against frequency. The effects of various speaker loadings can be observed and also the effects of long connecting leads.

With a signal generator it can be used, as was described in Chapter II, for making overall acoustic response tests of a receiver, being then in operation as a modulator of variable frequency, or as a means of accurately tuning a R.F. carrier a few hundred or thousand cycles away from a mid-frequency for bandpass and adjacent channel tests.

CHAPTER VIII

VALVE TESTERS

A VALVE tester is a device that gives sufficient indication of the all-round goodness of a valve for an observer to ascertain whether the valve is going to perform satisfactorily in a wireless set or amplifier. For such purposes the goodness of a valve is dependent upon several factors, of which the chief are (*a*) the emissivity of the cathode; (*b*) the degree of vacuum inside the glass envelope; (*c*) in the case of indirectly-heated cathode valves used in mains sets, the insulation between heater and cathode, particularly when hot, as in working conditions; and (*d*) the mechanical condition of the electrodes inside the glass envelope, including their insulation from each other.

It is perhaps best to consider each of these factors separately at first, in order to perceive how, in commercial valve testers, ingenious arrangements enable their effects to be observed by the movement of a pointer across a meter dial. Taking the last-mentioned factor first—the mechanical condition and insulation of the electrodes—the testing of this factor is very simple. A valve with mechanically defective electrodes will, save in rare cases, be "microphonic"—that is, it will set up anything from crackles when tapped to continuous howling as soon as the set is switched on. Modern manufacturing processes have gone a long way towards making electrode structures strong and rigid, and towards attaching the various parts of the structure to each other very firmly—usually by spot welding—but perfection is impossible; a turn of the control grid may become detached, or a strut of the anode, and hang loose or ready to vibrate with the slightest jar; overheating, even of short duration, of the filament of a battery valve may cause it to sag without breaking and become vibrant when jarred. There is no quantitative test possible, or needed, for valve microphony—the ear is a sufficiently competent tester, if the valve is lightly tapped.

Inter-electrode Insulation

The insulation between the electrodes of a valve is best tested by some sort of megger or neon circuit, since it must be extremely high to be satisfactory. Several commercial valve testers have something of the sort incorporated, for the preliminary test of a valve in this respect before its insertion in the tester, since a short-circuit between electrodes might damage the tester.

Taking the matter of cathode-heater insulation next, this is a factor that can be measured quite accurately. It is essential, however, that it should be measured "hot" as well as "cold"; and, if possible, with working voltages applied to the other electrodes. This is because leakage paths between heater

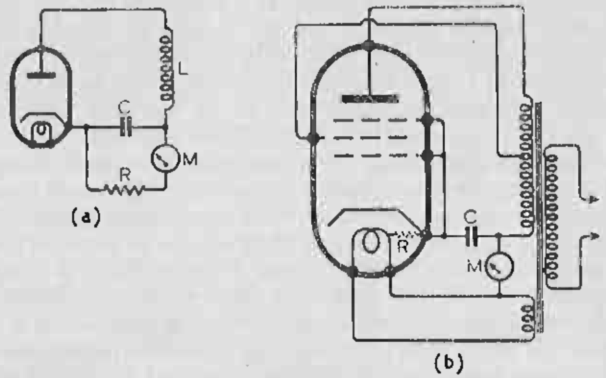


FIG. 37

(a) Indirectly-heated diode "leak and condenser" rectifier. (b) Circuit showing how the same principles are used to indicate faulty cathode-heater insulation in a valve when the cathode is hot, as in normal operation

and cathode quite often do not appear until the normal working heat has expanded the parts involved. Tested when cold, the insulation between cathode and heater may appear quite satisfactory—i.e. some 1 MΩ at least, and usually very much higher; as soon as the valve is put back into the receiver and used, it may give rise to intolerable hum. It would, of course, be possible to test hot cathode-heater insulation directly with a fairly obvious arrangement insulating the heating source very carefully from the cathode. But a rather simpler arrangement from a practical viewpoint, especially when mains supplies are used for power, as is usual in valve testers, is given in Fig. 37.

How it works may be understood by reference to Fig. 37 (*a*), which is that of a simple diode with a leak and condenser.

When an alternating voltage is developed across the coil L, the condenser C becomes charged on account of the rectifying action of the diode and discharges through the leak resistance R, which normally is included deliberately in such a circuit as this. A sensitive microammeter in series with R, as shown by M in Fig. 37 (*a*), would measure the value of the discharge current. A similar principle is involved in the circuit of Fig. 37 (*b*). Here, the required alternating voltage is supplied by the secondary of a mains input transformer (of which another secondary provides the heater supply in addition) and the leak resistance R is provided by the leakage path, if any, between cathode and heater, as shown in dotted line. If insulation is good, i.e. R is very high, little, if any, movement will be observed of the pointer of the meter M, the dial of which may even be calibrated, if desired, in terms of megohms leakage. The valve under test—in the circuit shown it is a pentode—is being made to act as a rectifier.

Softness

The vacuum inside the envelopes of modern valves is extremely high, except in the case of special types such as certain power rectifiers, neon types, etc. The slightest crack in the envelope, even if invisible, or defect in the sealing between the glass and the leads-out of the electrodes, lets in sufficient air to ruin this vacuum. Moreover, the heating of the electrodes when the valve is in use may cause them to release a residuum of occluded gas. Elaborate precautions are taken against it, of course, in manufacture, but valves are still liable to "go soft" in this way.

The easiest way to test a valve for softness is to measure the anode current with and without a high resistance between the control grid and the cathode. If the vacuum is satisfactory, the presence or absence of high resistance (say 2 MΩ) between grid and cathode should make no appreciable difference to the anode current. If the valve is soft, however, the anode current will not be the same in both cases: with high resistance in the grid circuit, the anode current will be greater than when the grid is returned directly to cathode.

Finally, there is the factor of emissivity of the cathode which, apart from anything else, determines inevitably the life of a valve. Most modern receiving valves are expected, if

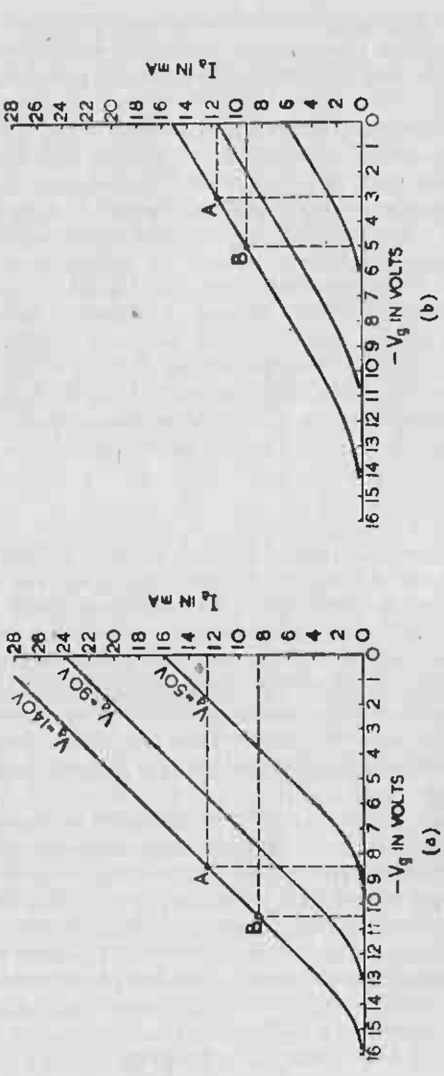


Fig. 38

(a) and (b) are I_a/V_a characteristics with normal and low cathode emissivity. Note the change in the slope of the lines AB in each case

used normally, to have a satisfactory operational life of about 1500 hours, although a lot depends on how much distortion or other falling off in performance a user is willing to tolerate. The important effect of loss of emissivity lies in the change it makes in the slopes of the families of characteristic curves. Two main families of curves are outlined in Figs. 38 (*a*) and (*b*), and the thing to note about them is that a change in cathode emissivity makes a change in the slope of the lines to a greater extent than does a change of any of the other parameters.

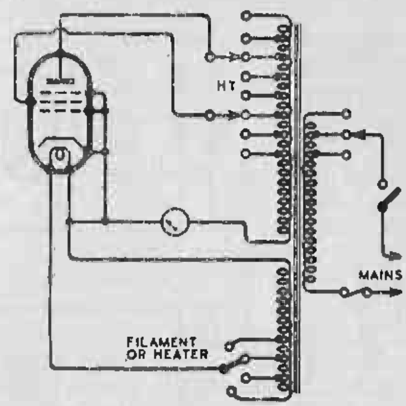


Fig. 39

Circuit showing a simple form of cathode emission valve tester, using a.c. only

Now, the constants of valves μ, R_a, and G_m are derived from the slopes of the characteristics; hence change in the emissivity of the cathode affects the valve constants considerably.

A few commercial valve testers merely test the emissivity of the cathode, without attempting to give any idea of the slope of any characteristic. It is assumed that if a valve does not pass the normal anode current when the filament or heater current, and the voltages applied to the other electrodes, are normal, the slopes of the characteristics will be altered. As a rough check on the condition of a valve, this is often good enough.

Instruments of this class are comparatively simple in construction, require only meters of a very moderate order of sensitivity, and are easy to use. When mains driven, they consist of little more than a transformer with multi-tapped secondaries for filament or heater supplier and for h.t. voltages to the other electrodes, a table being given for switching to the correct taps for the various types of valves to be tested. The valve being a unidirectional current device, the registering meter is merely a d.c. moving-coil instrument. A skeleton circuit showing the principles is given in Fig. 39. The scale of the meter is usually marked arbitrarily and is used with a table of the deflections to be expected with various types of valve.

Another form of valve tester does reveal approximately the slope of the I_a/V_g characteristic, by showing the amount of

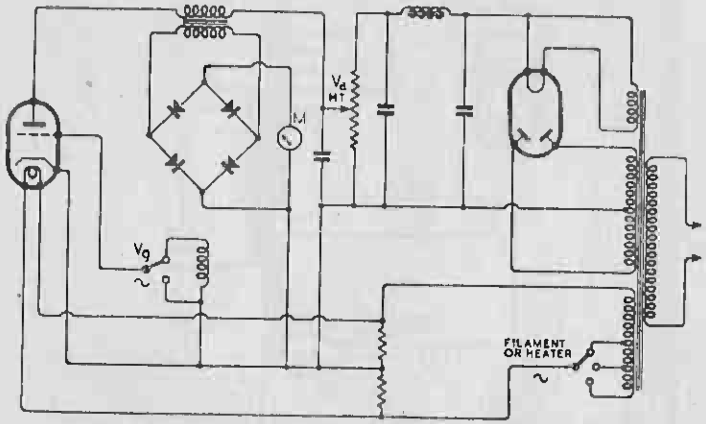


FIG. 40

Simplified circuit of a form of valve tester that registers the effect, in a rectifier type a.c. meter, of the a.c. produced in the anode circuit of a valve by the application of a suitable alternating voltage to the grid

alternating current component developed in the anode circuit of a valve when a standard alternating voltage is applied to the grid. The principle is shown by the circuit given in Fig. 40. It will be evident from a consideration of the I_a/V_g characteristic shown in Fig. 38 (*a*) that the application of an alternating voltage, to the grid of a valve having this characteristic, of a valve of about 0·75 volt R.M.S. will produce an a.c. component superimposed on the steady d.c. flowing in the anode circuit (with 140 volts as V_a) of about 2 mA R.M.S. If, however, the slope is less steep, as in Fig. 38 (*b*), the same V_g will produce only about 1·5 mA.

To make the point clearer, it will be assumed first, referring to Fig. 40 in conjunction with Fig. 38 (*a*), that the grid of the valve is not excited from the voltage developed across the coil L_1, but is kept steady at $-$ 9·5 volts, while V_a is 140 volts. A *steady* current will flow through the primary of the meter transformer, at 10·5 mA. Since this current is perfectly constant, no voltage is induced across the secondary of the transformer and the meter will, therefore, register zero. Now if the grid is connected to a tap on coil L_1 such that the potential on the grid rises to $-$ 8·5 volts, returns to $-$ 9·5 volts, falls to $-$ 10·5 volts, and returns again to $-$ 9·5 volts—as would happen if the voltage developed by this tap on L_1 developed a sine-wave alternating voltage of 1 volt peak (0·707 volt R.M.S.)—the current through the transformer primary would no longer be steady; it would start at 10·5 mA, rise with the grid voltage to 12·5 mA, return to 10·5 mA, fall to 8·5 mA, and return again to 10·5 mA. That is, an alternating component of a peak-to-peak value of 4 mA would appear in the current through the transformer primary, and this component would induce a voltage across the secondary that would operate the meter *M*. Applying the same alternating voltage to the grid of a valve of which the I_a/V_g characteristic had a less steep slope, as in Fig. 38 (*b*), would produce an alternating component of less than 3 mA with corresponding smaller effect on the meter.

In practice, of course, a valve tester of this type is mains driven, and L_1 consists conveniently of a winding on the mains power transformer, developing 0·5 volt R.M.S.—i.e. a peak-to-peak grid voltage swing of 1·4 volts—and the resulting alternating current component occurring in the anode circuit, varying in amount according to the slopes of the valves being tested, is registered on the rectifier-type meter *M*. Since manufacturers' valve data regarding mutual conductance (which is the slope of the I_a/V_g characteristic) are normally given with mean $V_g = 0$, no standing bias need be provided. Tables are provided to enable the testing conditions for each type of valve to be set up by use of control knobs and switches, and the meter dial may be marked in terms of mutual conductance.

Another type of valve tester measures the slope of the I_a/V_g characteristic directly—i.e. the actual change in anode current brought about by a change of grid potential of 1 volt is measured. Ingenious use is made of alternating voltage alone, without the complications of providing d.c. for both anode and grid voltages. Before considering the simplified circuit of this type, given in Fig. 42, the main principle can be grasped by a study of the circuit given in Fig. 41, in which d.c. is supposed to be employed. The grid of the triode valve that is supposed to be under test is taken to a bias source that

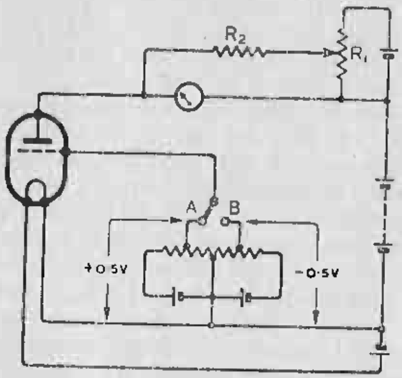


Fig. 41

Circuit showing principle employed in measuring the mutual conductance of a valve with d.c. power supplies

can provide a grid potential relative to the filament of either + or − 0·5 volt. When the switch connects the grid to 0·5 volt negative, the anode current is a certain I_1 as registered on the meter. If the grid is then switched to 0·5 volt positive —a total change of 1 volt—the anode current will change to I_2. Then the mutual conductance in terms of mA/V is $(I_2 - I_1)$ mA per volt.

Backing-off Circuit

This simple arrangement, apart from anything else, has one serious snag: if the valve under test is, say, a power output type passing a standing I_a of the order of 20 or 30 mA, the small *change* of I_a—which is the important factor—will not be easily or accurately observed on a meter that is capable of passing such a heavy current. To enable a sensitive meter to be used, therefore, which will be capable of giving a large readable deflection with or change of I_a of less than 1 mA, a backing-off circuit is provided, formed by a cell, in series with

the positive end of the h.t. battery, with a potential divider, R_1, across it. The slider of this potential divider is adjusted, with the valve passing anode current when the grid is switched to 0·5 volt negative, so that the voltage developed across the meter (through R_2) from this backing-off circuit is exactly equal and opposite to the p.d. set up across the meter itself

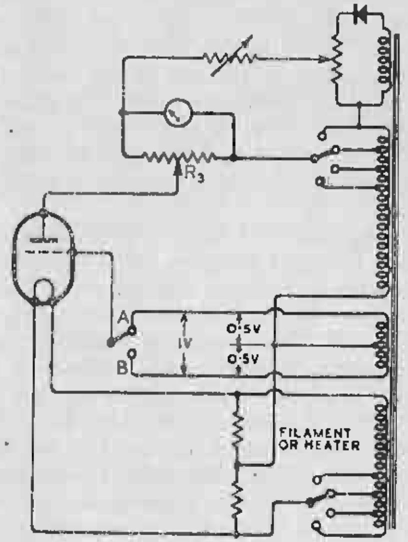


Fig. 42

Simplified circuit of a valve tester using alternating supplies to provide the working voltages for measurement of mutual conductance

by the passage of anode current through it. The meter then gives no deflection. As soon, however, as the grid is switched to 0·5 volt positive, the *increase* of anode current is registered on the meter, since the p.d. across the meter due to this increase of current becomes greater, whereas the voltage due to the backing-off potential divider remains the same as before. Thus, even a change of a fraction of a milliamp of I_a is readily observed on a sensitive meter without it being affected by the passage of heavy current. The resistance R_2 should be many times greater than the resistance of the meter in order to avoid an excessive shunting effect.

From the foregoing, the operation of the circuit of Fig. 42 can be derived. Referring to this figure the anode voltage is obtained from the h.t. secondary winding suitably tapped on the mains input transformer, as is the heater or filament supply from another tapped winding. The valve, therefore, only passes anode current during the positive half-cycles of voltage at the anode end of the secondary winding. How much anode current it passes depends on whether the grid is connected by the switch to the "in-phase" or the "out-of-phase" side of the centre-tapped winding (additional to the two previously mentioned) that supplies grid potential. "In-phase" and "out-of phase" refer to the polarity of the grid potential in relation to the anode potential; i.e. if the grid is positive when the anode end of the h.t. winding is positive, it is "in-phase," and vice versa.

Assuming, then, that the valve is in position and that suitable anode and filament voltages have been selected by the switches, the procedure is on the following lines: first, the grid is switched to the "out-of-phase" end of the grid voltage winding, say at B; the resulting standing anode current (actually, of course, unidirectional pulses of current each positive half-cycle at the anode end of the h.t. secondary) is backed off by the adjustment of the slider of a potential divider across a continuation of the h.t. winding until the meter registers zero. Then the grid is switched to the "in-phase" end of the grid voltage winding at A; this changes the effective grid potential by 1 volt in the positive direction, and the resulting increase of anode current is registered on the meter. The function of the half-wave rectifier in the backing-off circuit is to prevent a reverse current flowing through the meter during the negative half-cycles in the h.t. and backing-off windings.

The shunt potential divider R_3 across the meter may serve two purposes. One is to adjust the sensitivity of the meter circuit to suit various mutual conductance constants. The other may be to provide means of showing the state of a valve in a rough but attractive way by colourings or similar devices on the meter scale. The slider of this shunt potential divider is adjusted, by reference to a table or dial markings, to such a position that, if the increase of current brought about by the switching of the grid from the "out-of-phase" to the "in-phase" connection is normal, the meter is sensitive enough to

give nearly a full-scale deflection—or, for instance, for the pointer to swing over to a part of the meter scale marked "Good" or coloured brightly. If, however, the increase of current is below normal—indicating a less steep I_a/V_g characteristic, probably due to loss of cathode emissivity—the pointer only swings as far as the "Indifferent" or "Bad" part of the meter scale, or an appropriate colour which much impresses the lay customer, in whose view "meters do not lie"! Plainly, much depends on the accuracy with which the slider on the meter shunt is adjusted.

For the rest, rapidity and ease of preparation for test of many valve testers are facilitated by various forms of semi-automatic switching by means of perforated cards, numbered switches, etc., that also tend to make such instruments almost fool-proof. When a number of valves, of many different types, have to be quickly tested, such devices are of considerable and quite satisfactory service.

CHAPTER IX

MULTIVIBRATORS

THE multivibrator is a device that generates alternating potentials of non-sine waveform. The value of such a waveform lies in the fact that it can be analysed into a large number of harmonics of the fundamental frequency, and these harmonics can be selected by tuned circuits. Thus, the multivibrator may be regarded as a kind of generator at the output of which there are simultaneously present many frequencies, each an exact multiple of the fundamental. For example, if the fundamental is 100 kc/s, or 0·1 Mc/s, there may be available for use at the output as many as 500 frequencies, starting with 0·1 Mc/s and going up, at intervals of 0·1 Mc/s—i.e. 0·2, 0·3, 0·4, 0·5 Mc/s, etc.—to as high as 50 Mc/s. Whether they become usable at these high frequencies depends on the design and power of the multivibrator. At the other end of the frequency gamut, it is possible to arrange a multivibrator circuit to oscillate at a frequency as low as 1 c/s.

Further, it is possible to lock the fundamental frequency of a multivibrator readily to that of some high-precision generator, so that the multivibrator fundamental and all the harmonics each have precision of frequency of as high an order as that of the primary controlling frequency. One obvious use of such a controlled multivibrator lies in the quick and accurate calibration of receivers, especially when many of them are being so dealt with. The multivibrator is set going at, say, 100 kc/s, or 0·1 Mc/s, and its output fed to the input terminal of the receiver under tests. Then by tuning the receiver over the range required to be calibrated—say 6 to 8 Mc/s—signals will be found at every 0·1 Mc/s interval, and these points will give a fairly accurate calibration scale. When once set up, the multivibrator does not have to be touched or retuned; all the required frequencies for calibration are present simultaneously in its output, and each has the same accuracy as that of the primary controlling or triggering frequency.

The multivibrator is a comparatively simple device to construct and may be either battery or mains driven. It consists essentially of two resistance-capacitance coupled

amplifiers arranged in a mutual feed-back circuit as shown to the right of the dotted line in Fig. 43, which is a practical circuit as it stands. Such an arrangement—quite apart from the crystal oscillator on the left of the dotted line, which is the accessory triggering source—will oscillate on its own in the manner peculiar to this type of circuit; it is, in fact, a form of "relaxation" oscillator.

Frequency Control

The action of the multivibrator has been fully described by a number of writers, and the reader is referred to their

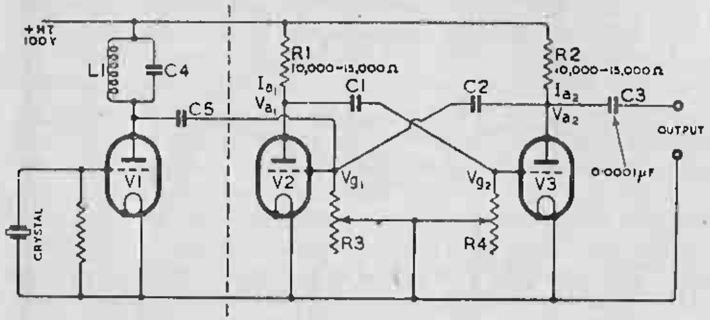


FIG. 43

The multivibrator circuit is shown to the right of the dotted line after C_5. The values of C_1, C_2, R_3, and R_4 depend on the frequency required, and are calculated from formulae given in the text

detailed works for the rather long explanations necessary to give full information concerning the precise action. Briefly, the action is such that the anode and the grid potentials of the two valves V_2 and V_3 go alternately positive and negative in relation to each other, rather like see-saws; when the potential of V_2 anode is up, that of V_3 anode is down, and vice versa, and when the grid potential of V_2 is negative, that of the grid of V_3 is positive, and vice versa. The frequency with which the reversals or alternations of potential take place is determined mainly by the rate of discharge of the coupling condensers C_1 and C_2 through the grid leaks R_3 and R_4; large condenser capacitances and high-value grid leaks give rise to low-frequency alternations and small capacities and low ohmic values of grid leak cause high-frequency alternations. Moreover, the shape of the wave of potential in time is very far from being sine; hence, the output taken off either grid or anode can be analysed into a large number of harmonically related component sine waves that can be selected by suitable tuned circuits.

Left entirely to itself, the multivibrator circuit does not perform its reversals at a very constant frequency, since at the moments of equilibrium, any random voltage, however tiny, can set off the change-over of conditions a little before or

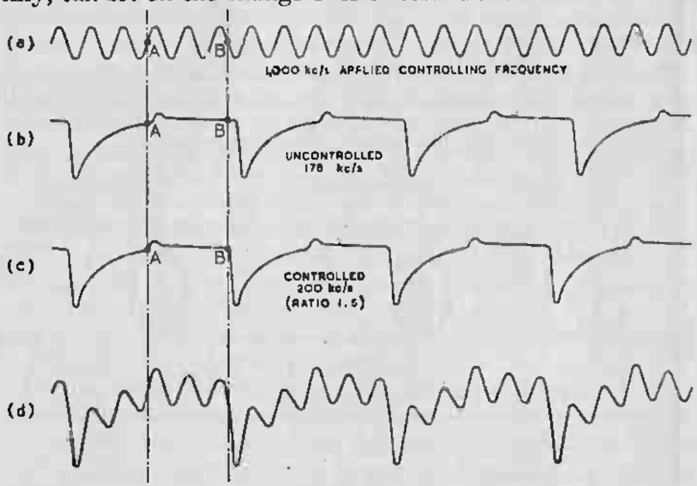


FIG. 44

Diagram showing the action of a triggering sine-wave potential in controlling the frequency of a multivibrator. (*a*) is the controlling sine-wave voltage. (*b*) is the uncontrolled multivibrator grid potential. (*c*) is the grid potential when controlled—without the sine-wave fluctuations superimposed. (*d*) is the resultant that would appear on the screen of a cathode-ray oscilloscope

after it would occur naturally. This leads on to the use of a controlling or triggering voltage that may be injected from an external generator of high precision as regards frequency. A glance at the waveforms shown in Fig. 44 will indicate the outlines of the effect of applying a controlling voltage to a grid.

An uncontrolled waveform is shown in Fig. 44 (*b*). On this line, at point *A*, the potential of the grid is not very far from the "zero" or change-over value—i.e. the value of negative potential at which anode current just begins to flow. It can be easily jolted, as it were, towards the zero line by an injected voltage in the right phase, which is at point *A* in Fig. 44 (*a*).

If it is large enough, this controlling voltage can cause the change-over to take place a little earlier than it would do naturally if solely dependent on the circuit constants of the multivibrator alone.

Once the change-over is started, the grid and anode potentials very soon move well away from the zero line and so can remain unaffected by the comparatively small controlling voltages. This brings out the interesting and valuable fact that a natural multivibrator frequency can be controlled by a voltage of sine-wave form at a much higher frequency—in practice, as much as ten times the frequency. The net effect as shown on an oscilloscope is a wavy line of the shape shown in Fig. 44 (*d*). This waveform is obviously composed of a very large number of harmonics as well as the fundamental. Whether these harmonic voltages are of sufficient amplitude to be usable depends partly on the power in a.c. form developed by the multivibrator, partly on the order of harmonic, and partly on the overall range of frequencies included by the fundamental and its harmonics.

In a usual type of battery-driven vibrator used for R.F. work at what are conventionally taken to be the "short waves," up to the 100th harmonic of a 200 kc/s fundamental is normally found of practical utility. Beyond that, not only are the harmonics weak, but they are so closely adjacent that it is difficult to distinguish between them.

In this connection, it will be useful to note how the particular harmonic in use can be ascertained when working amongst the high ranges. If it is desired, for example, to ascertain whether one is working on, say, the 34th or the 35th harmonic, the procedure is to tune a C.W. receiver exactly to any harmonic of that order (using, for instance, the zero-beat method), and then change the receiver tuning to exactly double the frequency. If this cannot be done with the receiver scale alone, a simple temporary oscillator can be set up to beat with the multivibrator harmonic and the second harmonic of the oscillator used to indicate twice the frequency. While tuning the receiver through to this second oscillator harmonic, the whistle obtained as the *multivibrator* harmonics are passed is counted. Then, when the receiver tuning is halted dead on the oscillator second harmonic (which gives the last whistle) it will then be tuned to the multivibrator harmonic given by 2 × *N*, where *N* is the number of whistles that have been counted. For example, let it be supposed that a receiver covers 18 whistles in moving from some one multivibrator harmonic to exactly double the frequency at which the receiver was first tuned. Then it will end up on the 2 × 18th or 36th harmonic. With this as a reference point, other harmonics can be distinguished by counting whistles up and down the tuning range. Further, if the fundamental is known the precise frequency involved at any harmonic can be calculated. Thus, if the multivibrator fundamental is 200 kc/s, the 36th harmonic would occur at a frequency of 200 × 36 kc/s, or 7·2 Mc/s.

It will have been noticed in Fig. 44 that it is necessary, if locking is to take place, for a not excessive difference in phase between the controlling and the multivibrator frequencies to be present. When widely asynchronous, either locking does not take place or it may do so at some undesired ratio between the controlling and natural frequencies. Hence, it is necessary that the multivibrator should be designed so that its natural uncontrolled frequency is *approximately* some desired ratio in respect to the controlling frequency chosen.

Frequency Calculations

The natural frequency of a multivibrator is determined mainly by the time-constants of the coupling condensers and grid resistances—C_1, C_2, R_3, and R_4, in Fig. 43—and when some particular multivibrator fundamental is desired, it can be derived approximately from the formula

$$f = \frac{1}{R_3 C_1 + R_4 C_2}$$

where R_3 and R_4 are in megohms, C_1 and C_2 are in microfarads, and f is in cycles per sec. If it is assumed that the time constants are to be equal so that $(R_3C_1) = (R_4C_2)$ the values of the capacitances and resistances can be found as a first rough design approximation from

$$R = \frac{1}{2fC} \text{ and } C = \frac{1}{2fR}$$

These are only approximations to give an idea of the order of resistance or capacitance required, and in making up a multivibrator, it is usual to make the grid resistances variable, the two being ganged on the one controlling shaft. They can then be adjusted in use until locking takes place.

The upper limit of oscillation frequency is set by the

amplification obtainable with resistance-capacitance coupling. If the high-order harmonics are to be preserved at reasonable strength, the whole circuit must be designed as for R.F. amplification with low losses.

The multivibrator can also be used for the production of audio frequencies and may be triggered by some stable and accurate A.F. generator such as a magneto-striction bar or

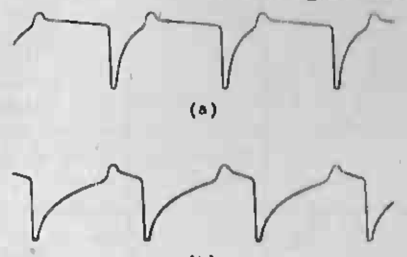
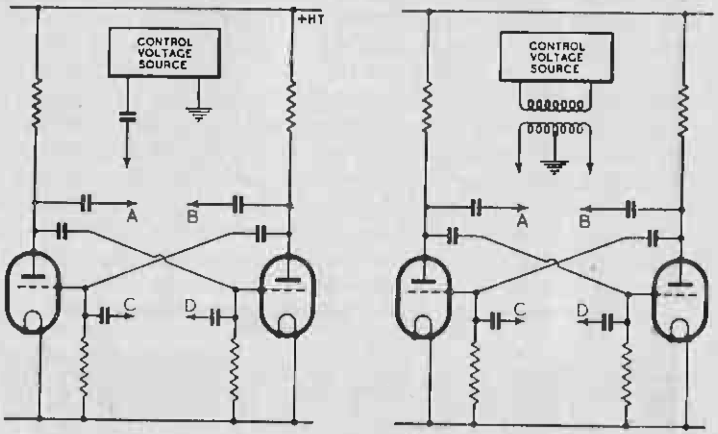


FIG. 45

Diagram showing the effects of differing time constants on the grid potential waveform

electrically driven tuning fork; the main difficulties then arise in designing sharply tuned A.F. filters for the selection of the desired harmonics in the output, which is not so easy at A.F. as it is at R.F. It is usual to employ tuned bridges for the purpose, or else to start with a comparatively high controlling A.F. (say 10 kc/s) and use a series of vibrators to obtain lower frequencies in multiples of ten. Usually the time constants are made equal in the two amplifiers. If they are unequal, the effect is to make a difference between the duration of the positive and negative part-cycles of grid or anode potential, as shown in Fig. 45. In the upper waveform (*a*), the negative part-cycle is of shorter duration than the positive. This indicates that the time constant of the condenser and resistance connected to the grid at which the potential is being observed is larger than that of the other. If the potential were observed at the other grid, it would be found to have the shape shown in the lower waveform (*b*).

The ratios of the controlling to the natural frequency at which locking tends to take place are determined by factors that are not so much abstruse as rather laborious to study. They will become clearer if several extended representations are made, to show the phase relationships between the natural and controlling frequencies at grids and anodes. In the first place, the ratios obtainable can only be integers, or complete numbers. In the second place, this fact limits the range of frequencies at which a multivibrator can be set up for locking with a given controlling frequency. For example, if the controlling frequency is 1000 kc/s (a crystal oscillator, for instance)



(a) (b)

CIRCUIT	CONTROL APPLIED	RATIOS CONDUCIVE TO OSCILLATION
(a)	A, B, C OR D ONLY	ODD OR EVEN
(a)	A & B SIMULTANEOUSLY OR C & D SIMULTANEOUSLY	EVEN
(b)	A & B IN "PUSH-PULL" OR C & D IN "PUSH-PULL"	ODD

FIG. 46

Table showing the ratios between the controlling and multivibrator fundamental frequencies to be expected with a few common circuit arrangements

possible ratios are only 5, 8, and 10. At a lower ratio than 5 the multivibrator frequency would be too high to be easily set up; at the higher ratios, only 5, 8, and 10 are exactly divisible into 1000. Hence, in such a multivibrator, locking could take place only when the fundamental was 200, 125, and 100 kc/s. Which of these possible ratios was selected would depend partly on the relative magnitudes of the controlling and natural potentials, and partly on where the controlling voltage was injected—whether at grid or anode, singly or at both.

The effect of increasing the controlling voltage is to make the multivibrator jump from one locking frequency to the next lower, with increasing uncertainty and instability as the divergence between the natural and the controlling frequencies increases. The effect of the point of application of the controlling frequency is perhaps set out in the simplest practical form of a table referring to Fig. 46, so that whether the output of the controlling generator is applied to one grid or one anode or to two grids or two anodes depends on whether odd or even ratios are desired.

The multivibrator lends itself to an increasing number of practical uses. It has been mentioned earlier as a source of precision signals in a laboratory or workshop for calibration or frequency determination purposes. It is also employed in various types of electronic switching circuit. One of its most interesting uses lies in very high accuracy primary or sub-standard wave- or frequency-meters in which the general arrangement is to control a series of multivibrators by a single extremely high-precision crystal oscillator. The generating of frequencies starts with this temperature-controlled crystal, which is placed in an oven regulated by a gas-discharge valve circuit and a bridge that is unbalanced by changes of temperature. The output of this crystal oscillator is fed, through buffer amplifiers, partly to an output terminal and partly to a multivibrator working at a fundamental 1/10th of the crystal frequency (which is usually 1 Mc/s). The output of this multivibrator is fed partly to an output terminal and partly to another multivibrator which is working at a fundamental of 1/10th that of the first vibrator. This goes on, in a chain of multivibrators, until one at a frequency of 100 c/s ends the gamut of frequencies covered. The output of the 1000 c/s stage is amplified sufficiently to work an electric clock, the time shown by which to a small fraction of a second may be checked by star transit observations with astronomical instruments or against time signals from some standard transmission. With such costly and elaborate apparatus, an accuracy better than one part in many millions is obtainable with a constancy extending over many months. In fact, frequency can be known to a much higher accuracy than either the velocity of transmission or wavelength.

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