# MEASUREMENTS

IN

# RADIO ENGINEERING

BY

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# PREFACE

The aim of "Measurements in Radio Engineering" is to provide a comprehensive engineering discussion of the measuring problems commonly encountered by radio engineers. The method of treatment and the degree of difficulty are much the same as in the author's book "Radio Engineering," and the two works are in a sense companion volumes in which one deals with the general principles of radio, while the other is devoted to measuring methods and measuring apparatus.

The present volume is intended to be of use as both a reference and a text book. For the practicing engineer it gathers together information on measuring techniques and measuring equipment, and thereby will be found of assistance when new problems are to be attacked. To the student it presents in an organized and systematic form a complete picture of the laboratory methods and laboratory measuring equipment ordinarily used in radio and allied fields. This makes it possible for the student to study the experimental aspect of radio in the same comprehensive way that the general principles are ordinarily studied. In particular, the present volume will be found of especial value in connection with laboratory courses where emphasis is placed on training in laboratory and experimental methods. Its use under such circumstances insures a broader viewpoint and better perspective than can be gained solely from a series of individual experiments, and also assists in the treatment of topics upon which experiments are not feasible.

Considerable attention has been given to the principles involved in the design and construction of laboratory equipment; partly because of the desirability of understanding thoroughly the equipment which one is using, and partly because of the fact that many circumstances arise where it is necessary to build one's own equipment in order to have any equipment at all. The latter situation arises particularly in university laboratories, where it is usually possible to construct valuable apparatus with the aid of student help. There are also many occasions in which special measuring equipment not commercially available is needed, and must be designed and built by the user.

"Measurements in Radio Engineering" has been written with the idea of presenting an engineering treatment of the subject, and is in no sense an encyclopedia of measuring methods. Success in making measurements in radio work is primarily a matter of having available a satisfactory technique that is thoroughly understood rather than of having available innumerable alternatives. Emphasis has therefore been placed upon those methods which experience has shown to be the most practical, which require the minimum of equipment, and which are least likely of error. Where alternatives are given, an attempt has been made to weigh their relative merits.

In view of the fact that in making radio measurements one is to a considerable extent utilizing expedients and ingenious practical applications of general principles, the techniques employed are much more subject to change than are fundamental principles. The information available in the literature on laboratory methods and the details of laboratory equipment is scattered and in many cases incomplete. The author is therefore fully aware that there is undoubtedly information which it would have been desirable to include in this book but which is known only to limited groups. Any suggestions which readers may care to send in for incorporation in later editions will be greatly appreciated.

Little emphasis has been placed upon detailed laboratory experiments suitable for use in university courses devoted to radio. This is because there are potentially many more experiments available than there is ever time to perform, and the particular ones which are selected depend to some extent upon the preference of the individual instructor and to a much greater extent upon the apparatus available.

Emphasis has chiefly been placed on the types of measurements that can be carried out within the limitations of a university laboratory course. It is believed that the main value of this type of book in connection with a university laboratory course is not as a manual of experiments, but rather as a textbook on general measuring principles which can be used to supplement the specific experiments carried out in the laboratory.

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STANFORD UNIVERSITY, CALIFORNIA,
*July*, 1935.

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# MEASUREMENTS IN RADIO ENGINEERING

## CHAPTER I

## VOLTAGE, CURRENT, AND POWER

**1. Measurement of Direct Voltage and Current.**—Direct currents of the magnitudes usually encountered in communication work are measured with portable instruments of the moving-coil (D'Arsonval) type. Such instruments are rugged, stable, and consume relatively little power. They are also commercially available in a wide variety of ranges, types, and accuracies.

Voltmeters are essentially current instruments provided with a series resistance. The power consumed by the voltmeter depends upon the current sensitivity of the instrument and is commonly expressed in "ohms per volt." Thus if 2 ma is required to give full-scale deflection, the required series resistance is 500 times the voltage which is to give full-scale deflection. The sensitivities commonly used are in the range 100 to 1333 ohms per volt, with the higher resistances usually preferred for communication work.

Meters to be used for laboratory and experimental purposes can be protected against accidental overloads by quick-acting fuses. It is also often desirable to provide shunts or multipliers to extend the range of an instrument. When this is done, scales should be arranged so that each division represents one, two, or five units. Hence when 50 divisions are available, the various ranges should be chosen out of the sequence 1—2.5—5—10—25, etc.; and when 100 divisions are available, the ranges should be taken from the sequence 1—2—5—10—20, etc.

*Multirange Voltmeters.*—Circuit arrangements for voltmeter multipliers are shown in Fig. 1. Both of the arrangements shown give equally satisfactory results, although the one at *a* is somewhat more economical in the use of resistances. The switch is preferably a tap switch and should be of the short-circuiting type; *i.e.*, as the blade passes from one contact to the next, it should momentarily short-circuit the two contacts rather than open the circuit. The resistances must be so chosen that at each position of the tap switch the total resistance in the circuit, including fuse and meter, has the proper value. For high-precision instruments the multiplier resistances should be made of wire having zero temperature coefficient (such as manganin) and should operate with a relatively low temperature rise. However, when the guaranteed accuracy of the instrument is of the order of 1 or 2 per cent, practically any type of wire-wound resistance unit is suitable provided it is operated at considerably below the usual wattage rating.

*Multirange Current Meters.*—In attempting to provide a current instrument with a number of ranges, care must be taken

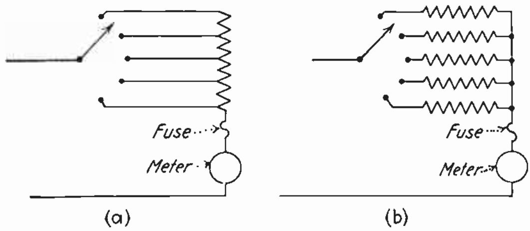



FIG. 1.—Voltmeter multiplier circuits.

to arrange the circuit so that contact and fuse resistances are not included in the shunt. Suitable circuit arrangements are shown in Fig. 2. Each arrangement requires a two-gang switch of the short-circuiting type. It will be noted that in both circuits the resistance of the fuse and the contact resistance of switch $S_1$ are in the line circuit and have no effect upon the multiplying ratio. The contact of the switch $S_2$ is in series with the much larger meter resistance, and so it introduces no trouble provided the switch is maintained in good condition. In the circuit shown at Fig. 2*a*, each shunt is adjusted individually to the appropriate range. Analysis of the simplified universal shunt circuit of Fig. 2*c* shows that the *relative* multiplying ratio is proportional to $R/R_1$ and is independent of both the meter resistance and also the total shunt resistance $R$. Hence if the resistance $R$ is tapped at points which make $R/R_1 = 1, 2, 5,$ and 10, then the *relative* multiplying factors at these same taps are 1, 2, 5, and 10, respectively, and are the same for any meter, whatever its resistance.

The total resistance $R$ of the universal shunt must be such that at position 2 in Fig. 2*b* the proper full-scale deflection will be obtained. An alternative arrangement is to make $R$ greater than this value, and then to connect across $R$ an additional shunt that is adjusted to give the desired sensitivity. The use of such an auxiliary shunt across $R$ does not affect the relative multiplying factors for the various taps and makes it possible to use the same shunt with instruments having slightly different resistivities.

Shunts of the type shown in Fig. 2*a* for currents up to 1 amp. can be wound on small spools turned out of doweling, and

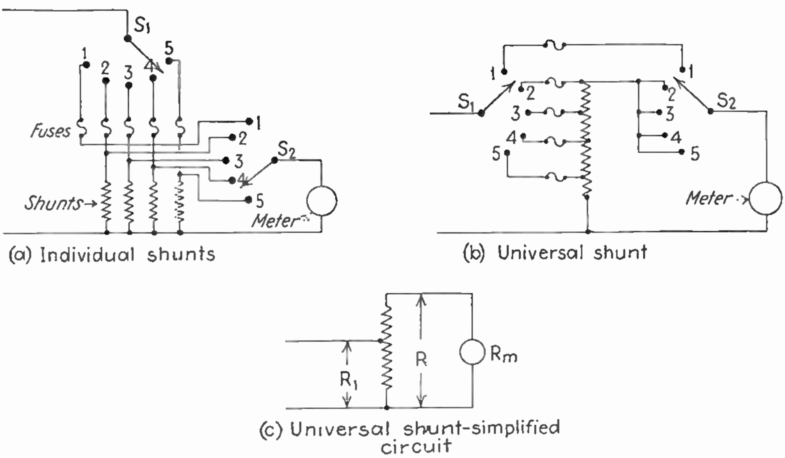



FIG. 2.—Circuits possible for current multipliers. The switches $S_1$ and $S_2$ should be of the short-circuiting type operated by a common shaft.

provided with an axial hole for mounting. The universal shunt can be more conveniently wound upon a card. Shunts can ordinarily be made of copper or any available resistance wire.[1] The only factor that must be kept in mind is that shunts of advance, nichrome, etc., will generate sufficient thermoelectric voltages at their terminals to cause small deflections on sensitive microammeters and milliammeters immediately after the shunt is soldered, or when in a warm place. When this effect is to be avoided, one should use either copper or manganin.

[1] Precision instruments having very low temperature coefficients require a special shunting arrangement. See "Standard Handbook for Electrical Engineers," 6th ed., p. 171, 1933.

**2. Calibration of Direct-current Instruments.**—The practical problem usually encountered in calibrating measuring instruments or adjusting multipliers is to use a single standard instrument having a single range, for checking both voltmeters and milliammeters of a great variety of ranges.

The arrangement shown in Fig. 3 is satisfactory for adjusting individual shunts and checking the calibration of milliammeters and microammeters. Here a standard voltmeter is used to provide a known voltage or a series of known voltages of the order of 10 volts or more. The resistance $R$ is so adjusted that this resistance, plus the resistance of the fuse, meter, etc., gives the desired current in the circuit when the known voltage is applied. The total resistance actually in the circuit at any time can be determined by throwing switch $S$ to the right and measur-

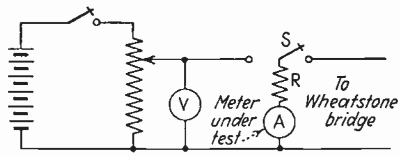



FIG. 3.—Convenient circuit arrangement for calibrating an ammeter or milliammeter $A$ using a standard voltmeter $V$.

ing on a Wheatstone bridge. The problem of adjusting the universal shunt is somewhat more complicated. One method of procedure is to wind the resistance $R$ of Fig. 2*b* to the appropriate value, and then find the positions of the various taps by trial, using a sharp-pointed prod to do the exploring. The sensitivity of current instruments can be most readily adjusted to a desired value by controlling the strength of the permanent magnet, using the technique and equipment employed by garages to adjust speedometer magnets. It is possible in this way to vary the sensitivity by as much as 10 per cent.

Voltmeters are most easily adjusted by first adjusting or determining the current sensitivity of the moving-coil instrument. The series resistance that is required can then be readily calculated and measured accurately by using a Wheatstone bridge.

The ultimate standard for all direct-current calibrations is the standard cell. The potential of such a cell is known to a very high order of precision and can be balanced against other potentials by means of a sensitive galvanometer and suitable potentiometer. The technique and apparatus for carrying out the necessary operations have become well standardized and are to be found in any work on instruments.

**3. Methods of Measuring Alternating Voltages and Currents.** The problem of measuring alternating currents in communication work is complicated by the wide frequency range which must be covered and by the resulting calibration difficulties. Although many methods have been devised for measuring alternating-current quantities, the only means used extensively in communication work are the iron-vane, dynamometer, rectifier, hot-wire, thermocouple, and vacuum-tube voltmeter instruments.

Iron-vane instruments are widely used for voltage and current measurements at 60 cycles. They are inexpensive, fairly accurate, and can be used up to frequencies in the order of 500 cycles without undue error. They can be made in maximum sensitivities of about 15 ma and hence a maximum ohms per volt of about 67. Dynamometer instruments are more accurate than iron-vane instruments but considerably more expensive. They are accurate up to frequencies of several hundred cycles to 1000 cycles depending upon the construction. Hot-wire instruments were once widely used for making current measurements at radio frequencies but are now virtually obsolete. They employ a small resistance wire tightly strung between two mountings. Upon heating, this wire stretches slightly and by mechanical linkage causes a pointer to indicate the amount of lengthening, and hence the current.

Rectifier instruments, thermocouples, and vacuum-tube voltmeters are the principal measuring instruments used at frequencies higher than 60 cycles and are described in detail in subsequent sections.

The behavior of alternating-current measuring instruments with non-sinusoidal waves of current is particularly important because these instruments are nearly always calibrated in terms of the effective value of the current, assuming a sine wave. Iron-vane, dynamometer, hot-wire, and thermocouple instruments give deflections that are exactly proportional to the effective value of the wave passing through them. Thus, if this wave consists of components of different frequencies having magnitudes $I_1$, $I_2$, $I_3$, etc., these instruments will give the same deflection as would a sine wave having an effective value of

$$\sqrt{I_1^2 + I_2^2 + I_3^2 + \cdots}.$$

Substitution in this formula shows that the presence of a 20 per cent second harmonic increases the reading by 2 per cent. The indication is not influenced by the relative phase positions of the various harmonics.

With rectifier and vacuum-tube voltmeter instruments the situation is somewhat different. The rectifier instrument gives a deflection which is proportional to the average value of the wave after rectification, and harmonics affect this average by an amount which depends not only upon the size of the harmonics but also upon their relative phase. With vacuum-tube voltmeters the behavior depends upon the adjustment, and it is possible to make the indications proportional to the effective value of the wave, to the effective value of the positive half cycle of the wave, to the average value of the positive half cycle, or to the peak amplitude. These different situations are discussed below and summarized in the table on page 21.

**4. Rectifier Instruments.**[1]—In the rectifier instrument the current to be measured is passed through a full-wave copper oxide rectifier unit and the resulting direct current indicated by a moving-coil direct-current instrument. Rectifier instruments can be built to give full-scale deflection on currents less than 1 ma and so make possible the construction of alternating-current voltmeters having sensitivities of the order of 1000 ohms per volt. The ruggedness and overload capacity compare favorably with moving-coil direct-current instruments. The best accuracy obtainable is, however, only about 5 per cent, because of the variation of rectifier characteristics with temperature. Furthermore, the electrostatic capacity of the rectifier element partially by-passes the rectifier at the higher audio frequencies, causing an additional error amounting to approximately ½ to 1 per cent for each thousand cycles. Another disadvantage of rectifier current instruments is that they have a high voltage drop, a 1-ma instrument consuming about 1 volt at full-scale deflection.

The characteristics of the rectifier instrument depend primarily upon the characteristics of the rectifier element. The most important of these are shown in Fig. 4 for low audio frequencies

[1] For further information see Joseph Sahagen, The Use of the Copper Oxide Rectifier for Instrument Purposes, *Proc. I.R.E.*, vol. 19, p. 233, February, 1931.

such as 60 cycles. It will be observed that the rectified d-c current is very nearly linearly proportional to the a-c current except at very small currents, where the direct current tends to be proportional to the square of the alternating current. The resistance of the rectifier unit depends very considerably upon the alternating current passing through the rectifier, being greatest when the current is small.

As a consequence of these characteristics an unshunted rectifier instrument measuring current will have an approximately linear scale provided the full-scale current is not too small. The same is true when a high series resistance is employed to give a

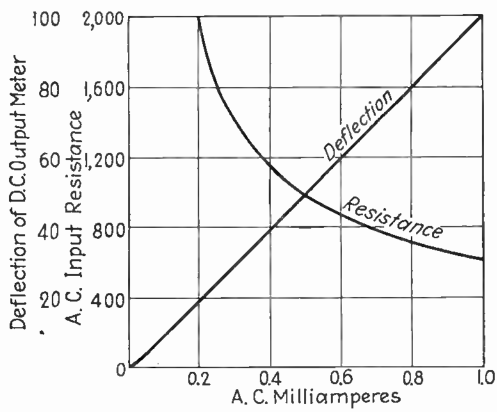



FIG. 4.—Rectified d-c current and resistance to alternating current expressed in terms of the alternating current delivered to a typical copper oxide rectifier type of instrument.

high-range voltmeter, but when a small series resistance is used for a low-range voltmeter the variable resistance of the rectifier element causes the current through the instrument to decrease faster than the voltage, so that the scale is bunched near the low end.

Rectifier instruments give an indication which is proportional to the average value of the alternating-current wave when the negative areas are treated as though they were positive. Since the ratio of average to effective values of a sine wave is 0.909, this means that the direct-current instrument used with the rectifier must have a sensitivity approximately 10 per cent greater than the alternating current to be measured. Because the rectifier instrument gives an indication proportional to the average value of the wave, it is susceptible to wave-form errors which, in general, will depend upon the phase as well as the magnitude of the harmonics, as summarized in the table on page 21.

*Multirange Rectifier Instruments.*—The problem of providing a rectifier meter with several ranges is complicated by the fact that the rectifier resistance varies with current. If the different ranges are to make use of a common scale, it can be shown that the equivalent resistance of the network connected across the rectifier terminals must be constant, irrespective of the multiplying factor which this network introduces.[1]

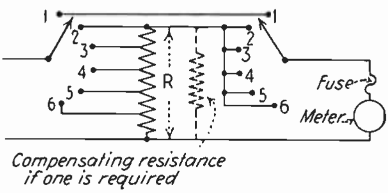



Fig. 5.—Multiplying network for rectifier milliammeter that follows a common scale for different ranges. The two switches shown are operated from a common shaft.

With current instruments this means that the resistance of the multiplier, as viewed by the rectifier when the source of current supply is opened, must be constant. The simplest arrangement satisfying this requirement is the universal shunt, which can be incorporated in a multirange current instrument as shown in Fig. 5. This is similar to the corresponding circuit for the direct-current case shown in Fig. 2*b* except that only a single fuse is now required because the resistance of the rectifier is very large compared with possible variations between fuses. When a multirange instrument employing the circuit of Fig. 5 is used, point 1 gives the full sensitivity of the unshunted instrument and, accordingly, requires a substantially linear scale as indicated by Fig. 4. The remaining points fit a common nonuniform scale, which will be more nearly linear the higher the shunt resistance in proportion to the instrument resistance.

Voltmeter multipliers employing a common scale must be so designed that the resistance which faces the rectifier when the source of voltage is short-circuited is constant. The simplest arrangement to accomplish this is shown schematically in Fig. 6, in which, as the series resistance $R_1$ is increased, the shunt

[1] For a more detailed discussion see F. E. Terman, Multirange Rectifier Instruments Having the Same Scale Graduation for All Ranges, *Proc. I.R.E.*, vol. 23, p. 234, March, 1935.

resistance $R_2$ is decreased. The formulas for this case are given below:[1]

$$R_1 = \frac{V}{I_m\left(1 + \frac{R_m}{R_{eq}}\right)} \tag{1a}$$

$$R_2 = \frac{R_1 R_{eq}}{R_1 - R_{eq}} \tag{1b}$$

where

$V$ = voltage which is to give full-scale deflection

$I_m$ = current which must be delivered to rectifier to give full-scale deflection

$R_m$ = resistance offered to alternating current by rectifier when the current $I_m$ is flowing through the rectifier (measured with meter and fuse in circuit)

$R_{eq}$ = resistance of multiplier as viewed by the rectifier instrument = $R_1R_2/(R_1 + R_2)$

$R_1$ = series resistance of multiplier

$R_2$ = shunt resistance of multiplier.

The scale will be more nearly linear the higher the ratio $R_{eq}/R_m$. The lowest voltmeter range determines the highest possible value of this ratio which, for greatest sensitivity, requires that the shunt resistance $R_2$ be infinite.

For high voltages, such as 50 volts or higher, it is possible to dispense with the arrangement shown in Fig. 6 and use a simple series resistance, since at these higher voltages this resistance is so much higher than the meter resistance that variations in the latter introduce negligible error. Under these conditions the calibration curve for voltage is the same as for current and is

[1] These equations are derived as follows: By reference to Fig. 6 one can write

$$\left.\begin{matrix}\text{Current through } R_1\\ \text{produced by } V\end{matrix}\right\} = \frac{V}{R_1 + (R_2R_m/(R_2 + R_m))}$$

The fraction $(R_2/R_m + R_2)$ of this current flows through the meter, so

$$I_m = \frac{VR_2}{[R_1 + (R_2R_m/(R_2 + R_m))](R_m + R_2)}$$

This can now be reduced to Eq. (1) by remembering that by definition

$$R_{eq} = \frac{R_1R_2}{R_1 + R_2}$$

accordingly substantially linear. The most satisfactory way of providing a rectifier voltmeter with a wide variety of ranges is hence to use the constant-resistance attenuator arrangement shown in Fig. 6 for the lower voltage ranges, and to use a simple series resistance for the higher ranges. Two scales must then be provided, one of which is nearly linear and is to be used when the series resistance is employed, while the other is a non-uniform scale graduated to fit the particular value of $R_{eq}/R_m$ being employed. The complete circuit diagram for such a voltmeter is shown in Fig. 7.

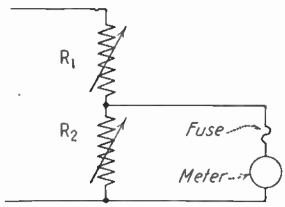



FIG. 6.—Schematic diagram of multiplier for copper oxide type of rectifier voltmeter that follows the same scale for different ranges.

The multiplier resistances used in rectifier instruments can be any of the commercial wire-wound radio resistors padded out to within 1 or 2 per cent of the proper value, or they can be made by winding resistance wire on flat cards. No special care need

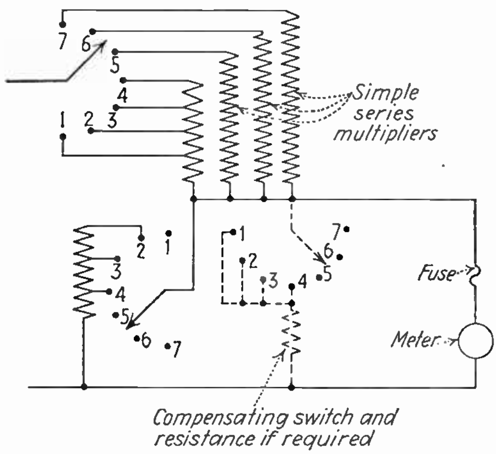
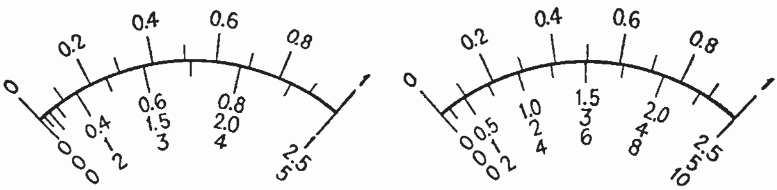



FIG. 7.—Circuit diagram for multiplier network that will enable different ranges of a rectifier voltmeter to follow the same scale graduations. The switches are all operated from a common shaft.

be taken to avoid capacity and inductance effects since the error thus introduced in multiplier impedance will in all cases be negligible compared with other errors of rectifier instruments.

*Example of Multirange Rectifier Instrument.*—As an illustration of the procedure to be followed in designing multiplier systems for rectifier instruments, consider the problem of providing a rectifier meter having the characteristics shown in Fig. 4 with current ranges of 1, 2.5, 5, 10, 25, and 50 ma. The 1.0-ma range is, of course, supplied by the unshunted instrument (position 1 in Fig. 5) and requires a linear scale graduation. The 2.5-ma range is obtained at position 2 of Fig. 5, and to give correct sensitivity the universal shunt must hence have a total resistance of two-thirds of the resistance $R_m$. Reference to Fig. 4 shows the meter resistance at full-scale deflection is 619 ohms, and, allowing 20 ohms fuse resistance, it is thus necessary that the total shunt resistance $R$ of Fig. 5 be $(619 + 20) \times \frac{2}{3} = 426$ ohms. The 5-, 10-, 25-, and 50-ma ranges represent multiplying factors of 2, 4, 10, and 20, with respect to the 2.5-ma



(a) Milliammeter Scale (b) Voltmeter Scale

FIG. 8.—Scale graduations that would be used in the multirange rectifier instruments worked out in the text.

range, so that these require taps on $R$ that are 213, 106.5, 42.6, and 21.3 ohms from the bottom of $R$ (*i.e.*, at 426 divided by 2, 4, 10, and 20, respectively).

The scale graduations corresponding to these shunted ranges are shown in Fig. 8*a*, together with the substantially linear scale for the 1-ma range. These scales will have to be hand drawn until commercial multirange rectifier instruments are made available.

Assume now that an instrument having the characteristics shown in Fig. 4 is to be made into a voltmeter having full-scale ranges of 2.5, 5, 10, 25, 50, 100, and 250 volts. The lowest four ranges will use the multiplier system of Fig. 6, while the higher ranges will use a simple series resistor. In order to keep the scale as linear as possible, the multiplier resistance $R_{eq}$ for the lower ranges will be made as high as possible. This is accomplished by making the shunt resistance $R_2$ equal to infinity for the most sensitive (2.5-volt) range. The series resistance $R_1$ for this condition is then 2500 − 639 = 1861 ohms, since for full-scale deflection with 2.5 volts the total circuit resistance must be 2.5/0.001, and all but the 639 ohms accounted for by rectifier and fuse must be supplied by $R_1$. The value of $R_{eq}$ also equals 1861 ohms, since by Eq. (1*b*) $R_1 = R_{eq}$ when $R_2 = \infty$. Substituting $R_{eq} = 1861$, $R_m = 639$, and $I_m = 0.001$ into Eqs. (1*a*) and (1*b*) shows that the values of $R_1$ and $R_2$ required for the various ranges are as given in the accompanying table. The 50-volt and higher ranges use simple series multiplier resistances of 1000 ohms for each volt, including the meter resistance of 639 ohms as part of the multiplier resistance.

TABLE I.—COMMON SCALE RECTIFIER VOLTMETER
$R_m = 639$ $I_m = 1.0$ $R_{eq} = 1861$

| Volts for full scale | $R_2$ | $R_1$ | Ohms per volt (full scale) | Scale used |
|---|---|---|---|---|
| 2.5 | ∞ | 1,861 | 1000 | Non-uniform |
| 5.0 | 3722 | 3,722 | 855 | Non-uniform |
| 10.0 | 2481 | 7,444 | 797 | Non-uniform |
| 25 | 2068 | 18,610 | 765 | Non-uniform |
| 50 | ∞ | 49,400 | 1000 | Linear |
| 100 | ∞ | 99,400 | 1000 | Linear |
| 250 | ∞ | 249,400 | 1000 | Linear |

The scale arrangements that should be used on the voltmeter are shown in Fig. 8*b*. The substantially linear scale is for the 50-, 100-, and 250-volt ranges, and the non-uniform scale for the lower ranges.

*Standard Multipliers for Unlike Instruments.*—Individual rectifier instruments requiring the same current for full-scale deflection will normally be found to have resistance characteristics similar in general character but differing in absolute magnitudes from each other. The necessity of calculating and constructing a special multiplier for each instrument can be avoided by making all multipliers alike and designing them to offer an equivalent resistance $R_{eq}$ which is slightly higher than the value required by the highest resistance instrument. This standard multiplier can then be made to fit the individual instruments by shunting the multiplier output with a resistance that is adjusted to bring the combined resistance across the instrument terminals down to the desired value. A compensating resistance of this character is shown dotted in Figs. 5 and 7. In the case of voltmeters it is necessary to add an extra switch in order that this shunt may be removed on the higher voltage ranges where the multiplier consists of a simple series resistance. The addition of the compensating resistance to a standard multiplier as shown in Figs. 5 and 7 does not in any way affect the multiplying factors introduced. The only disadvantage is that the use of the compensating resistance increases slightly the voltage drop of ammeters and reduces the ohms per volt of voltmeters by a small amount.

The linearity of the instrument scale depends only upon the ratio $R_{eq}/R_m$ and not upon whether the instrument is to be a voltmeter or ammeter. As a consequence it is possible to make a combined voltameter by designing both voltage and current multipliers to offer the same resistance $R_{eq}$, since then a common non-uniform scale can be employed. The scale graduations approach more closely to a linear law the higher the multiplier resistance compared with the meter resistance (*i.e.*, the higher $R_{eq}/R_m$).

Rectifier instruments may be conveniently calibrated at 60 cycles. The calibration will then hold accurately up to 1000 cycles and will drop off slowly at higher frequencies at the rate of perhaps ¾ per cent per 1000 cycles. In the commercial instrument shops the exact sensitivity for full-scale deflection is controlled by varying the sensitivity of the direct-current galvanometer until the desired alternating current gives full-scale deflection.

**5. Combination Alternating-current—Direct-current Instruments.**—Since the rectifier instrument includes a D'Arsonval-type meter, it is possible to use the latter independently of the rectifier, as a direct-current voltmeter or ammeter to cover any desired ranges. This makes it possible to read both direct and alternating voltages and currents on a single meter.

The best method of constructing such a universal instrument is to employ two scales. A linear scale is used for the measurement of direct currents and voltages, for the most sensitive a-c current range (when no shunt is used), and for the higher alternating-current voltage ranges above about 50 volts. The remaining a-c current and voltage ranges make use of a common, non-uniform scale of the type described in the preceding section. In order that the current and voltage ranges may both fit the same non-uniform scale, it is merely necessary that the equivalent resistance $R_{eq}$ of the multiplier be the same for current and voltage cases.

It will be noted that since an effective a-c current of 1 ma will deliver approximately 0.909 ma to the direct-current instrument, a galvanometer suitable for use in a 1-ma rectifier instrument will also, with suitable shunts, be capable of measuring d-c currents from 1 ma upward and will give a sensitivity in excess of 1000 ohms per volt when used as a voltmeter.

**6. Thermocouple Instruments.**—In a thermocouple instrument the current to be measured heats a short piece of resistance wire that is associated with a thermocouple. The output of the

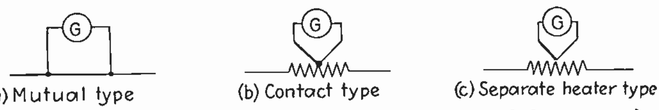



FIG. 9.—Schematic diagram showing different types of thermocouple instruments.

thermocouple is recorded by a sensitive direct-current microammeter which thus gives an indication of the a-c current passing through the heater wire. It is evident that, under conditions where there is negligible skin effect in the heater, there will be no frequency effect, and calibrations can be made with direct current. Because of this, thermocouple instruments are the standard means of measuring currents at radio frequencies and are also used at audio frequencies when accuracy is important.

Thermocouple instruments may be of several types. The more sensitive ones make use of a thermocouple and heater inclosed in a vacuum in order to reduce conduction heat losses. Instruments of this type are made for measuring currents from about 1 ma to 1 amp. For currents of 100 ma and upward, it is entirely satisfactory to operate the thermocouple heater in the air. The thermocouple may be of the mutual-, contact-, or separate-heater type, shown respectively at *a*, *b*, and *c*, in Fig. 9. In the mutual type the thermocouple serves as its own heater, but this has the disadvantage that the direct-current galvanometer shunts the heater. The contact type overcomes this disadvantage by connecting the thermocouple to the heater at only one point, but it is less sensitive than the mutual type because the thermocouple wires conduct heat away from the junction. In the separate-heater type the thermocouple is held near to but insulated from the heater by a small glass bead. This makes the instrument sluggish and slightly less sensitive than the contact type because of the temperature gradient in the glass bead, but it avoids the possibility of error that may arise in directly connected types as a result of electrostatic capacity between galvanometer and ground.

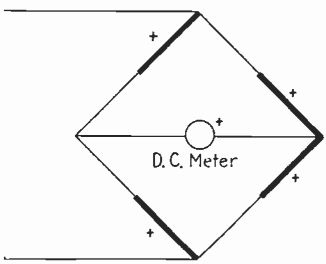


FIG. 10.—Bridge type of thermocouple instrument.

Commercial thermocouple instruments commonly employ the bridge circuit shown in Fig. 10. This arrangement preserves the high sensitivity of the mutual-type couple and avoids its disadvantages by placing the galvanometer across the neutral arm of a bridge.

A variety of materials may be used to form the thermocouple, of which the most common combination is constantan (or advance) against copper or manganin. This junction gives a thermal e.m.f. of approximately 45 microvolts per degree centigrade, and this coefficient is constant up to about 230°C. Thermocouples of this type can be conveniently made by spot-welding or soldering the wires together.

The direct-current microammeter associated with a thermocouple must have high current sensitivity and low resistance because the thermal e.m.f. is small. Thermocouples burn out with small overloads,[1] and satisfactory protection cannot be obtained by fuses since the normal operating condition of a thermocouple is very close to the burn-out point.

The deflection of thermocouple instruments is very nearly proportional to the square of the heater current, since when the heat is lost by conduction the temperature of the heater is proportional to $I^2R$, *i.e.*, to the square of the a-c current. In vacuum thermocouples operated near the burn-out point an appreciable

[1] Thus the rush of charging current that results when several hundred volts is applied to a 4 mfd condenser will invariably burn out the thermocouple of a 125-ma instrument even though the steady current flowing in the circuit is zero.

part of the heat is lost by radiation, and there is some departure from the square law. The fact that the heat generated in the heater is proportional to the square of the current accounts for the fact that the thermocouple instrument measures the effective value of a non-sinusoidal wave, as discussed in Sec. 3.

In calibrating contact- and mutual-type thermocouple instruments with direct current, it is always necessary to reverse the polarity of the d-c current and take the average reading for the two polarities as the true reading. This is because the resistance drop in the heater at the contact may cause a small amount of d-c current to flow through the microammeter; this effect must be averaged out by reversing the calibrating current. The calibration is reasonably permanent, although it must be checked periodically if precision is important.

For every thermocouple instrument there is a frequency above which the calibration depends upon frequency. This results from the fact that, if the frequency is high enough, skin effect in the thermocouple heater causes the effective heater resistance to increase, thereby increasing the heater temperature for a given current. The frequency at which this action first becomes appreciable is greater the smaller the heater wire, which leads to the result that thermocouples designed for small currents can be used at higher frequencies than those carrying large currents.

*Thermocouple Instruments for Measuring Large Currents.*—The measurement of large high-frequency currents introduces a number of difficulties. Thermocouple instruments with the heater carrying the entire current are not satisfactory because the heater would require a large wire that would then have considerable skin effect even at the lower radio frequencies. Ordinary shunts cannot be employed to carry the major part of the current since the shunting ratio will be affected by the relative inductances as well as resistances, thus introducing a considerable frequency effect.

One possible solution is an array of shunts arranged symmetrically as shown in Fig. 11*a*. Each filament of wire possesses equal inductance so that the current divides in the same way at high frequencies as with direct currents, provided the wires are uniform and have substantially identical resistances. The condenser shunt shown in Fig. 11*b* has also been used successfully in the measurement of large currents.[1] The current will divide between the two parallel condensers directly as their capacities and will maintain this ratio independent of frequency as long as the condenser that is in series with the thermocouple has a much higher impedance than the thermocouple heater.

Current transformers can be used to measure large radio-frequency currents with a low-range thermocouple instrument as shown in Fig. 11c. The fundamental principles involved in this instrument are the same as in current transformers for power

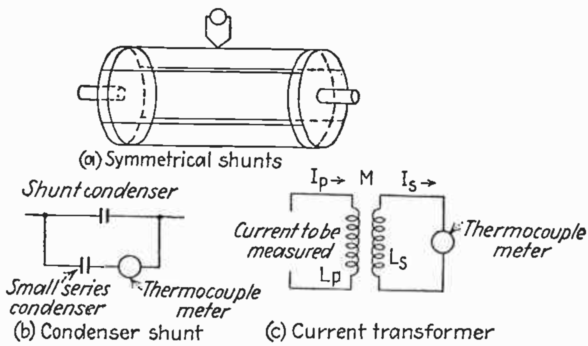



FIG. 11.—Methods that can be successfully employed to measure large radio-frequency currents.

frequencies. The factors controlling the transformation ratio are made clear by the following analysis. The voltage induced in the secondary by a primary current $I_p$ is $\omega M I_p$ where $M$ is the mutual inductance between primary and secondary and $\omega$ is $2\pi$ times the frequency. The resulting current which flows in the secondary is hence

$$\text{Secondary current} = \frac{\omega M I_p}{R_s + j\omega L_s}$$

The transformation ratio is hence

$$\frac{\text{Primary current}}{\text{Secondary current}} = \frac{R_s + j\omega L_s}{\omega M}$$
$$= \frac{L_s}{M}\sqrt{1 + (R_s/\omega L_s)^2} \qquad (2)$$

[1] See Alexander Nyman, Condenser Shunt for Measurement of High-frequency Currents of Large Magnitude, *Proc. I.R.E.*, vol. 16, p. 208, February, 1928.

If the total secondary circuit resistance $R_s$ is small compared with the inductive reactance of the secondary, which can be readily accomplished, the transformation ratio is seen to be independent of frequency.

Radio-frequency current transformers are commonly constructed with the secondary wound on a toroidal ring, through the center hole of which the wire carrying the primary current is looped once or twice. Current transformers for frequencies below 1,000,000 cycles often have the secondaries wound on a magnetic core.

**7. Vacuum-tube Voltmeters.**—The vacuum-tube voltmeter is essentially an ordinary detector tube in which the change in d-c plate current that takes place with the application of a signal

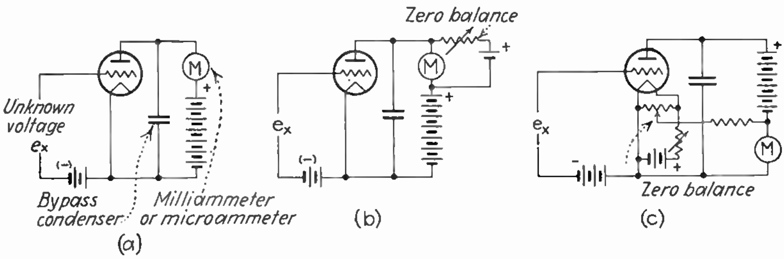



FIG. 12.—Typical vacuum-tube voltmeter-circuit diagrams. The arrangements at *b* and *c* provide means for balancing the normal current out of the meter *M*.

voltage is used to measure this voltage. When properly made, a vacuum-tube voltmeter can be calibrated at 60 cycles and used at all higher frequencies up to the highest encountered in radio work. The vacuum-tube voltmeter also consumes negligible power, thus having practically infinite ohms per volt, together with an input capacity which is in the order of 5 to 10 $\mu\mu$f. The voltage range most conveniently measured is from 1 volt to 10 or 20 volts, but this can be extended appreciably in either direction if necessary. As a result of these characteristics, the vacuum-tube voltmeter is probably the most important of all instruments for making measurements at radio frequencies.

Vacuum-tube voltmeters can be constructed in a variety of forms. The commonest arrangements are variations of the plate rectification circuits as shown in Fig. 12. In these arrangements the curvature of the plate-current grid-voltage characteristic of the tube causes the d-c plate current to change by an amount which is a measure of the applied voltage. This change in plate current is read upon the milliammeter in the plate circuit, which must be thoroughly by-passed to all alternating currents. It is convenient to balance out the steady plate current present when no signal is applied, so that the full range of the milli-

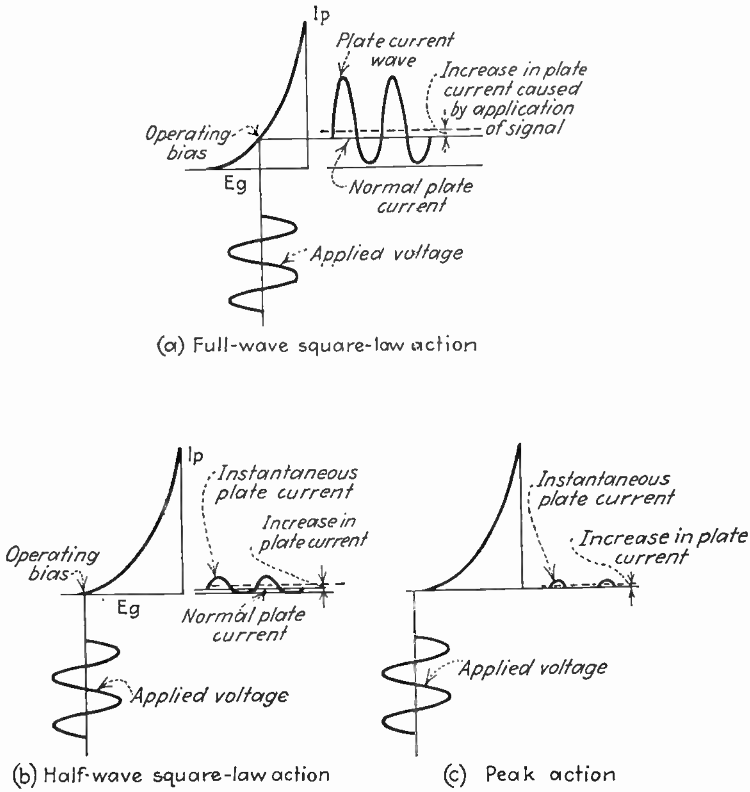



FIG. 13.—Diagrams illustrating action taking place in vacuum-tube voltmeters with operating points chosen to give full-wave square-law, half-wave square-law, and peak action.

ammeter in the plate circuit is available for reading change in plate current. Means for doing this are shown at *b* and *c* in Fig. 12.

*Characteristics of Vacuum-tube Voltmeters.*—The behavior of a vacuum-tube voltmeter depends to a considerable extent upon the part of the tube characteristic which is selected as the operating point. If the grid bias is so related to the plate voltage as to allow the plate current to flow continuously as at Fig. 13*a*, the change in plate current is very nearly proportional to the square of the effective value of the applied voltage. This is because the tube characteristic can be approximated by a parabola over a reasonable range. Under these conditions the vacuum-tube voltmeter is said to operate under full-wave square-law conditions. If the grid bias is so chosen that the operating point is substantially at cut-off, as shown in Fig. 13*b*, then the negative half cycles of the applied voltage are entirely suppressed and have no effect, while the change in plate current will be very nearly proportional to the effective value of the positive half cycles. This condition is termed half-wave square-law action. If now the grid bias is increased still more, as shown in Fig. 13*c*, the change in plate current is determined by the peaks of the positive half cycles, and the instrument tends to become a peak voltmeter.

When the vacuum-tube voltmeter is calibrated on sinusoidal voltages, and all voltages measured are sinusoidal, any of the adjustments, *i.e.*, full-wave square law, half-wave square law, or peak, will give the same results. However, if the voltage being measured contains harmonics, the vacuum-tube voltmeter reading will be influenced and the magnitude of the indication will depend upon the type of adjustment employed. In the case of the full-wave square-law adjustment, the instrument reads the effective value of the wave, and the result is independent of the phase of the harmonics. In the case of the half-wave square-law detector, the result depends not only upon the magnitudes of the harmonics but also upon their relative phase, and the same is true even to a greater extent with the peak adjustment. The effects to be expected are summarized in Table II.[1]

Reversing the input terminals of a vacuum-tube voltmeter will sometimes change the reading of the output meter. This is known as "turnover" and is caused by even harmonics in the applied voltage wave making the positive and negative half cycles differ in wave form. Turnover is greatest in peak-type voltmeters, is present in the half-wave square-law type, and non-existent in the full-wave square-law and linear types. When

[1] For further information on this subject, see Irving Wolff, Alternating Current Measuring Instruments as Discriminators against Harmonics, *Proc. I.R.E.*, vol. 19, p. 647, April, 1931.

turnover exists, the average of the direct and reverse polarity readings will be approximately but not necessarily exactly the correct value. The situation with respect to turnover is summarized in the accompanying table.

TABLE II

| | Full-wave square law | Half-wave square law | Full-wave linear[1] | Peak |
|---|---|---|---|---|
| Turnover possible? . . . | No | Yes | No | Yes |
| Phase of harmonics affect reading? . . . . . | No | Yes | Yes | Yes |
| Effect of harmonic component on reading. | | | | |
| 50 per cent second harmonic. | 11 per cent | −6 to +27 per cent | 0 to 10 per cent | −25 to +50 per cent |
| 50 per cent third harmonic. | 11 per cent | 12.5 per cent | −10 to +16 per cent | 8 to +50 per cent |

[1] This case corresponds to the copper oxide rectifier instrument.

The calibration curve of full-wave and half-wave square-law vacuum-tube voltmeters is very nearly a straight line when plotted on log-log paper, as shown in Fig. 14. This is a very

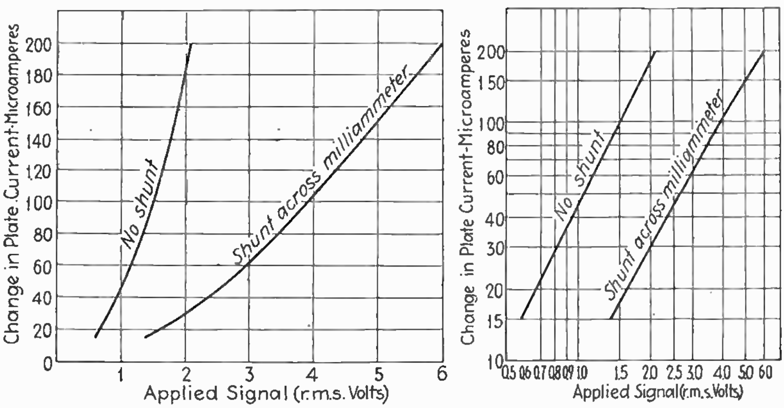



FIG. 14.—Typical calibration curve of vacuum-tube voltmeter operated as in FIG. 13*a* or Fig. 13*b*, when plotted on rectangular and logarithmic coordinates.

great convenience since it means that a complete calibration curve can be drawn with a relatively small number of points, and the calibration can be checked periodically with very little work.

*Design of Vacuum-tube Voltmeters.*—The design of vacuum-tube voltmeters for any given purpose can be reduced to a simple procedure. The first question is the choice of a tube. This is generally made on the basis of sensitivity, anode voltage required, and cathode heating power. The sensitivity of a vacuum-tube voltmeter is determined by the mutual conductance of the tube, while the plate voltage required is proportional to the amplification factor $\mu$ of the tube. As a result the best tubes for vacuum-tube voltmeter work are usually low-mu tubes with high mutual conductance, *i.e.*, power tubes.

The grid bias should be just slightly larger than the crest value of the largest voltage to be measured. The plate voltage is then chosen in relation to the bias voltage so that the operating point is on the desired part of the characteristic. Thus if the tube is to be operated at cut-off, the plate voltage will be $\mu$ times the grid bias, and hence $\mu$ times the crest value of the signal voltage expected. The by-pass condenser in the plate circuit should be of adequate size to act as a thorough short circuit at the lowest frequency to be measured, when compared with the plate resistance of the tube. The balancing system can be arranged in any convenient manner. When direct current is employed to heat the cathode, it is convenient to use the arrangement shown in Fig. 12*c* and so avoid the necessity of additional batteries.

All that now remains is to select the plate milliammeter and to provide it with the necessary shunts. The more sensitive the milliammeter, the smaller will be the voltage that can be read with accuracy. If the instrument gives a full-scale deflection on 200 microamperes it will be found that with ordinary tubes full-scale deflection will be obtained with an unknown voltage in the order of 2 volts effective. Larger voltages up to the point where the grid begins to draw current can be measured by providing the meter with suitable shunts.

*Miscellaneous Vacuum-tube Voltmeter Circuits.*—The simple circuit arrangements that have been shown for vacuum-tube voltmeters may be modified in many ways, some of which are shown in Fig. 15. The self-bias arrangement shown at *a* is commonly employed, and upon superficial analysis appears to have a number of advantages, since the fact that the grid bias will increase as the d-c plate current increases causes the calibration curve to be substantially linear. This linearity is gained at the expense of sensitivity, however. Furthermore, the fact that the bias increases when large voltages are applied causes the instrument to tend toward a peak device with large signals, thus introducing large errors on non-sinusoidal waves. The arrangement shown at Fig. 15*b* behaves similarly to the self-biased arrangement, having the same linear calibration and the same tendency toward being a peak voltmeter when measuring large voltages.

Another modification is shown at *c*. This is essentially a direct-current bridge in which effective plate resistance of the tube to direct current is one arm and the resistances $R_1$, $R_2$, $R_3$ are the remaining arms. The bridge is first brought into balance when the input is short-circuited. This is accomplished by varying the applied voltage with the rheostat $R$ until the galvanometer reads zero. It will be noted that this always brings both the grid and plate potentials to standard values, irrespective of deterioration of the anode battery. Application of the voltage to be measured will now upset this balance by changing the effective plate-filament resistance of the tube, thus causing current to pass through the galvanometer. The magnitude of this current is a measure of the unbalance and hence of the applied voltage. Such an arrangement has the advantage of automatically always bringing the voltage on the plate and grid of the tube to the same value, thus insuring greater permanence of calibration than would be otherwise possible.[1]

The design of a bridge-circuit type of vacuum-tube voltmeter as shown in Fig. 15*c* presents a number of special features. For ordinary use, the operating point should be taken as something less than cut-off because the bias increases when a signal is applied to the grid. By a suitable choice of the operating point it is possible to make the bias reach the cut-off value when the full signal voltage is applied. The resistance $R_3$ is chosen to achieve this, and the resistances $R_1$ and $R_2$ are assigned a ratio which is equal to the ratio of plate-filament to grid-bias voltage. In order to minimize the tendency for the grid bias to vary with signal voltage, the current drawn by $R_1$ and $R_2$ should be several times the d-c plate current when the full signal voltage is applied

[1] A bridge-type vacuum-tube voltmeter somewhat similar to this is described by S. C. Hoare, A New Thermionic Instrument, *Trans. A.I.E.E.*, vol. 46, p. 541, 1927.

to the grid. The sensitivity can be controlled by shunting the meter $M$. The battery voltage available must exceed the sum of grid-bias and plate voltages by perhaps 20 per cent in order to make possible the initial adjustment of balance.

The vacuum-tube voltmeter arrangement of Fig. 15$d$ employs the grid-leak method of detection instead of plate rectification. This is slightly more sensitive than plate rectification for very

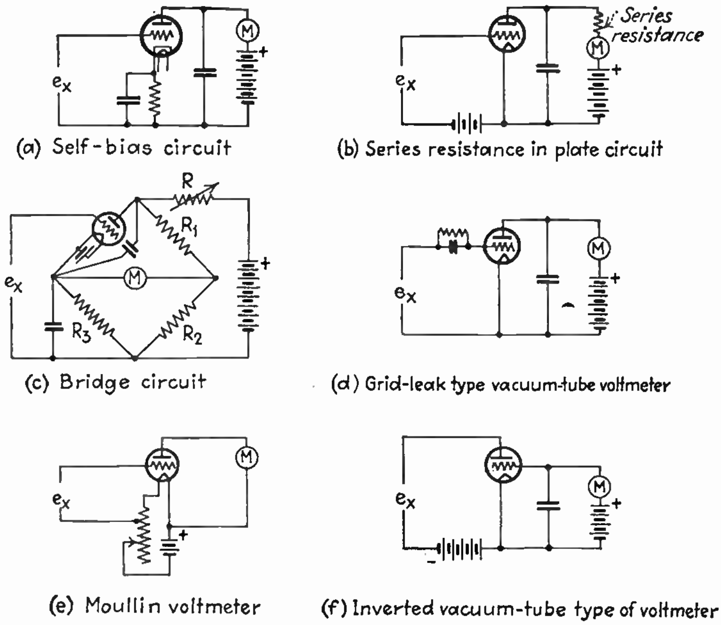



FIG. 15.—Miscellaneous vacuum-tube voltmeter circuits. These circuits may all be provided with means for balancing the residual current in the plate meter $M$ if this is desired.

small voltages but has the disadvantage that the residual plate current which must be balanced out is rather large.

The circuit at $e$ has been developed by Moullin[1] and is used extensively in England. It is characterized by an apparent lack of plate voltage, although actually the plate is slightly positive with respect to the main part of the filament because of the voltage drop in the filament.

Screen-grid tubes can be used as vacuum-tube voltmeters and have been recommended for such purposes by several investi-

[1] E. B. Moullin, "Radio Frequency Measurements," 2d ed., p. 140, 1931.

gators. They have no particular advantage from the point of view of sensitivity, however, and have the disadvantage of requiring an additional electrode voltage which is just one more factor to cause the calibration to vary.

When very large voltages are to be measured with a conventional vacuum-tube voltmeter, the grid bias required is extremely large and the plate voltage is correspondingly high even if the amplification factor is low. Under these conditions it is sometimes desirable to employ an inverted vacuum tube, in which the plate is used as the control electrode as shown in Fig. 15*f*.[1] If such a tube has an amplification factor of $\mu$ when used in the normal manner, it will have an amplification factor of approximately $1/\mu$ when inverted. Thus if a voltage of 200 is to be measured, the negative bias on the plate could be made $-200$, and the positive-grid voltage required for cut-off adjustment is only $200/\mu$.

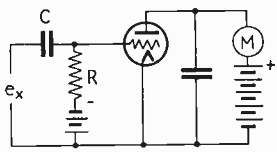



FIG. 16.—Grid-leak and grid-condenser arrangement for blocking direct-current potentials from the vacuum-tube voltmeter grid, and for always insuring that there is a conducting path by which the negative bias voltage can reach the grid.

*Use and Calibration of Vacuum-tube Voltmeters.*—In using vacuum-tube voltmeters, particular care must be taken to avoid impressing direct-current voltages upon the instrument along with the alternating-current voltage which is to be measured. This can be accomplished by the use of a suitable resistance-condenser combination such as shown in Fig. 16. This arrangement also insures that there will be a conducting path by which the negative grid-bias voltage can reach the grid. The blocking condenser $C$ must have a reactance less than one-tenth the resistance $R$ at the lowest frequency to be handled. The condenser leakage must also be very low, and mica dielectric is accordingly desirable.

A trouble commonly encountered in using vacuum-tube voltmeters comes from operating the voltmeter at a potential above ground. If the circuit being measured has a direct-current voltage to ground, it is essential that the vacuum-tube voltmeter be ungrounded. In order to avoid error, it is also desirable that the cathode side of the vacuum-tube voltmeter input be at

[1] See F. E. Terman, The Inverted Vacuum Tube, a Voltage-reducing Power Amplifier, *Proc. I.R.E.*, vol. 16, p. 447, April, 1928.

ground potential as far as alternating voltages are concerned. If this is not the case, errors and sometimes very peculiar behavior will result.

The calibration of a vacuum-tube voltmeter depends upon the grid, plate, and filament voltages and upon the tube characteristics. The tube is reasonably permanent, particularly if the filament is operated at slightly less than normal voltage; but unless some means such as shown in Fig. 15*c* is provided for checking and adjusting the grid and plate voltages to a standard value, frequent recalibration will be necessary. In using a vacuum-tube voltmeter provided with a balancing arrangement such as shown in Fig. 12*b* or 12*c*, it is necessary to check the accuracy of the zero balance at frequent intervals.

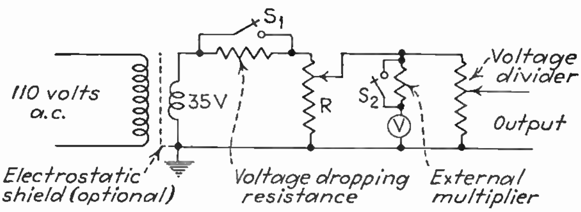



FIG. 17.—Apparatus for calibrating vacuum-tube voltmeter. Under normal conditions switch $S_1$ is open and $S_2$ closed, but by throwing both of these a multiplying factor can be introduced.

Since vacuum-tube voltmeters require relatively frequent calibration, it is convenient to provide equipment especially designed for this purpose. A suitable arrangement is shown in Fig. 17, in which a 60-cycle voltage is adjusted to a known value by means of a rheostat $R$ and voltmeter $V$ and then applied to a voltage divider which should have about 100 ohms per volt and be tapped at appropriate points to give convenient calibrating voltages. An electrostatic shield between transformer primary and secondary is desirable but not always essential.

For general purposes the meter $V$ can be an iron-vane instrument reading 15 volts full scale, with an external multiplier which will extend the range to 30 volts upon throwing a switch. The voltage dividing resistance may then be 1000 ohms tapped at 0, 10, 20, 40, 60, 80, 100, 125, 150, 200, 250, 300, 350, 400, 500, 600, 700, 800, 900, and 1000 ohms. This will provide voltages of 0, 0.1, 0.2, 0.4, 0.6, 0.8, 1.0, 1.25, 1.50, 2.0, 2.5, 3.0, 3.5, 4, 5, 6, 7, 8, 9, and 10, when 10 volts is applied to the divider, and twice these amounts when 20 volts is applied. Intermediate voltages can be obtained by making the applied voltage differ from exactly 10 or 20 volts.

*Peak Voltmeters.*—The peak value of a wave can be determined with the aid of the arrangement shown in Fig. 18*a*. The bias voltage as read upon the voltmeter *V* is adjusted by the potentiometer *P* until the milliammeter or microammeter *M* just shows signs of current. Under these conditions the voltage read by *V* is very nearly equal to the crest value of the alternating voltage applied across the input terminals. If the bias voltage is slightly less than the crest alternating-current voltage, the

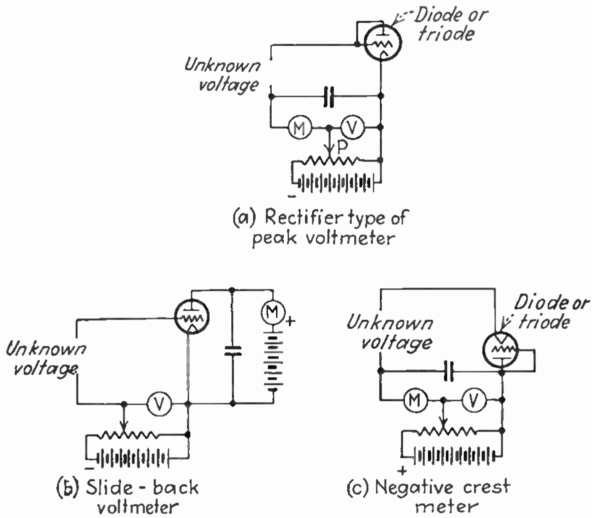



FIG. 18.—Different types of peak voltmeters.

anode will go positive around the peak of each cycle, causing current to flow; whereas, if the bias is the least bit greater than the peak alternating voltage, the anode is never positive with respect to the cathode and no current can possibly flow. Thus the voltage of *V* at which a trace of current flows represents to a high degree of accuracy the peak value of the unknown voltage. The voltage *V* is not, however, exactly equal to the peak of the unknown voltage because the velocity with which the electrons are emitted from the cathode will cause a few microamperes of current to flow even when the anode is never more positive than the cathode. This uncertainty is in the order of 1 volt and for very precise results it is necessary that the instrument be calibrated by determining the relationship between the peak value of the applied voltage and the voltage $V$ that exists when some small but definite current, such as 10 $\mu$a, is indicated by the milliammeter.

Another type of peak voltmeter, known as the "slide-back" vacuum-tube voltmeter, is shown in Fig. 18*b*. To use this instrument the input terminals are first short-circuited and the grid bias $V$ adjusted until the meter $M$ records a small plate current, say 100 $\mu$a. The unknown voltage is then applied and the bias increased until the plate current is again reduced to the same value. The increase of the bias is then very nearly equal to the peak value of the unknown voltage, although for precise results a calibration curve must be prepared and used.

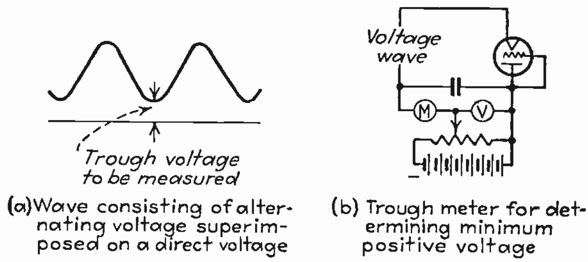



FIG. 19.—Trough meter for reading trough of a pulsating positive wave. Note that the battery polarity has been reversed from that in Fig. 18*c*.

The arrangements shown in Figs. 18*a* and 18*b* are arranged to measure the peak value of the positive half cycles. If the peak amplitude of the negative half cycles is desired, it is necessary to rearrange the circuit of Fig. 18*a* as shown at Fig. 18*c*.

There are certain circumstances where it is necessary to measure trough values of a wave. These occur when an alternating voltage is superimposed upon a direct-current potential of larger amplitude, as is the case in Fig. 19*a*. The minimum amplitude of such a wave can be determined by a negative crest meter such as shown in Fig. 18*c*, connected as shown in Fig. 19. The method of operation is similar to that of other peak voltmeters already described.

The peak voltmeters that have been described all have the disadvantage of requiring manipulation to obtain a reading. Direct-reading peak voltmeters can be built by using one of the circuit arrangements shown in Fig. 20. These are either grid-leak or diode detectors with a high-resistance grid leak and a relatively high-capacity grid condenser. In such a circuit the grid condenser is charged to within a few per cent of the peak voltage being measured. In the circuit at *a* the direct current which this condenser discharges through the leak as measured by the microammeter *M* will be proportional to the average voltage across the condenser. At *b* the voltage built up across the condenser affects the d-c plate current, which is read upon meter *M*. It will be noted that these instruments shown in Fig. 20 are direct reading in the same sense as any voltmeter or ammeter, and can be used for monitoring the operation of oscillators, modulators, audio systems, etc.

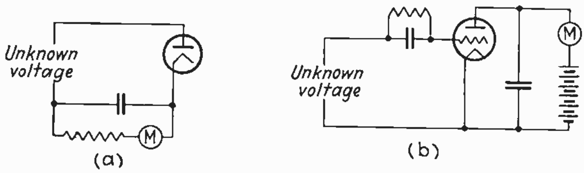
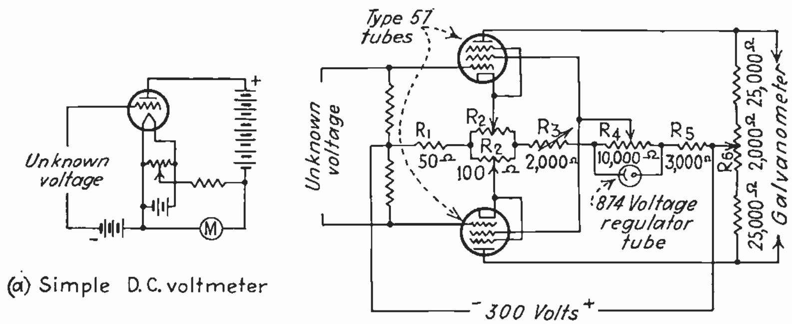



FIG. 20.—Direct-reading types of peak voltmeters.

*Amplifier-detector Voltmeters.*—Very small alternating voltages can be measured by placing an amplifier between the small potential to be measured and the vacuum-tube voltmeter (or detector). Such arrangements have wide use in the audio-frequency measuring technique employed in connection with telephone work. It is desirable that amplifier-detector voltmeters be used in conjunction with an audio-frequency standard signal generator (see Sec. 46). The general idea is to apply the unknown voltage to the input of the amplifier-detector and adjust the amplification until the output meter gives a convenient reading. The input terminals of the amplifier-detector are then shifted to a standard signal generator giving voltages of the same frequency, and the standard voltage is adjusted until the output is the same as before. The standard voltage is now obviously equal to the unknown voltage.

*Vacuum-tube Voltmeter Arrangements for Measuring Direct-current Voltages.*—Vacuum-tube voltmeters can be used to measure direct-current potentials while drawing substantially zero current. Suitable circuit arrangements are shown in Fig. 21. The circuit shown at *a* is an ordinary vacuum-tube voltmeter with the input voltage directly coupled to the grid, and is suitable for measuring potentials of a few hundred microvolts or more. When smaller potentials are to be measured, this simple arrangement gives trouble because slow drifts in the battery potentials obscure the very small changes in plate current which the unknown voltage produces.

To avoid this difficulty, numerous circuits have been devised in which changes in battery voltages are prevented from affecting the output galvanometer. A typical circuit of this type is shown in Fig. 21*b*.[1] This circuit is so proportioned that, if the supply voltage increases, the resulting increase in grid bias is just sufficient to neutralize the effect of the increased screen and




(a) Simple D.C. voltmeter

b Balanced D.C. voltmeter

FIG. 21.—Vacuum-tube voltmeters for measuring direct-current voltages.

anode potentials as far as the plate current is concerned. The two tubes in the balanced circuit are furthermore adjusted to balance out what residual variations are present.

The procedure for setting up the circuit is as follows: A meter is placed in the plate circuit of one tube and the resistance $R_3$ and slider on $R_4$ varied in such a way as to maintain substantially constant voltage upon the screen but with a varying fraction of this voltage obtained from $R_3$. This is continued until an adjustment is found for which the plate current is independent of the supply voltage. The meter is now transferred to the plate circuit of the second tube and the point of cathode return of this tube is adjusted until a similar balance is obtained. A galvanometer (usually a wall galvanometer) is now connected across

[1] For further details of this and of other similar circuits, see G. P. Harnwell and S. N. VanVorhes, A Balanced Electrometer Tube and Amplifying Circuit for Small Direct Currents, *Rev. Sci. Inst.*, vol. 5, p. 244, July, 1934.

the output terminals and the final adjustment made by varying the point of cathode return of one of the tubes while readjusting $R_6$ to maintain zero deflection in the galvanometer. It is possible to find a condition for which the galvanometer will not deflect when there are appreciable changes in the supply voltage.

In some cases where direct-current voltages are to be measured, it is essential that the tube draw the very smallest grid current possible. Special tubes are commonly employed for this purpose. These are very carefully evacuated and are designed to operate on electrode voltages in the order of 6 to 10 volts in order to make ionization of residual gas by collision impossible and to prevent X-rays produced by the impact of electrons at the plate from ionizing the residual gas.

**8. Power Measurements at Audio and Radio Frequencies.** The commonest method of determining power at audio and radio frequencies is to measure the effective resistance of the circuit and the r.m.s. current flowing. The power can then be calculated from the relation $P = I^2R$. This method will give satisfactory results at all frequencies since circuit resistance can be measured with good accuracy.

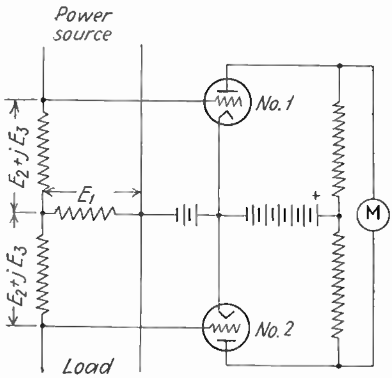



FIG. 22.—Circuit diagram of vacuum-tube wattmeter. The two tubes must have identical characteristics and be operated as full-wave square-law vacuum-tube voltmeters.

Power can also be determined in a number of other ways. The ordinary dynamometer wattmeter commonly employed at 60 cycles can be designed to operate at frequencies of several hundred cycles to 1000 cycles, although some accuracy is lost by inability to maintain the proper phase relationship between currents flowing in the voltage and current coils.

A pair of identical square-law devices can also be used to indicate power, as shown in Fig. 22 for the case of full-wave square-law vacuum-tube voltmeters.[1] Here the voltage applied

[1] See H. M. Turner and F. T. McNamara, An Electron Tube Wattmeter and Voltmeter and a Phase-shifting Bridge, *Proc. I.R.E.*, vol. 18, p. 1743, October, 1930; also U. S. Patent No. 1,586,553, E. O. Peterson.

to tube 1 is $(E_1 + E_2 + jE_3)$ where $E_1$ is proportional to the voltage of the circuit, $E_2$ proportional to the in-phase component of current that flows, and $E_3$ to the quadrature current, while the voltage acting on the grid of tube 2 is $(E_1 - E_2 - jE_3)$. By operating each vacuum tube so that it functions as a square-law device, the change in d-c plate current of the two voltmeter tubes is respectively proportional to $(E_1 + E_2)^2 + E_3^2$ and $(E_1 - E_2)^2 + E_3^2$. The difference between these is measured on the meter $M$ of Fig. 22 and is proportional to $E_1E_2$, which in turn represents power consumed by the load since it is the product of the voltage and the in-phase component of current. The principal limitation of this method is the difficulty of obtaining and maintaining absolutely identical characteristics in the two vacuum-tube voltmeters.

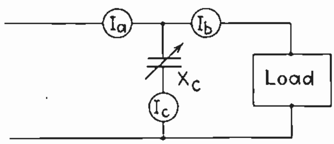



FIG. 23.—Three-ammeter shunt-reactance method of measuring power.

The three-ammeter shunt-impedance method of measuring power shown in Fig. 23 is often used to determine power in radio-frequency transmission lines, or the power developed by a radio transmitter. Analysis of this circuit shows that the power in watts flowing through the device is

$$\text{Watts} = 2X_c\sqrt{S(S - I_a)(S - I_b)(S - I_c)} \qquad (3)$$

where

$X_c$ = the reactance of the shunt condenser in ohms

$I_a, I_b, I_c$ = the currents in the three meters (see Fig. 23)

$$S = \frac{I_a + I_b + I_c}{2}$$

The best accuracy is obtained when the reactance of the shunt condenser is such that $I_b$ and $I_c$ are of the same order of magnitude. The accuracy becomes small when the load is highly reactive.

Photometric methods are sometimes used to determine radio-frequency power consumed in a load. The idea is to use an incandescent lamp as the load and then determine the direct-current power that is required to produce the same luminous intensity in the lamp as developed by the high-frequency energy.

The heat energy liberated in a circuit in a given length of time, and hence the power, can always be measured with a calorimeter. The difficulty here is that the apparatus is not generally available, the determinations take a long time, and, unless the losses are appreciable, the accuracy will be poor.

**9. Power Output and Power Level.** *Output Meters.*—An output meter is a device used for measuring the output power of amplifiers, radio receivers, etc., and, as ordinarily made, consists of a load resistance with which is associated a rectifier-type instrument calibrated directly in watts or decibels. Typical circuit diagrams are shown in Fig. 24. The load resistance may have taps so that it can be adjusted to match different power sources.

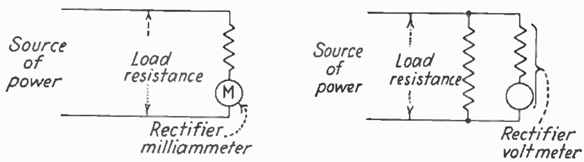



FIG. 24.—Simple types of output meters, where the load consists of the rectifier meter and associated resistances, while the rectifier meter is calibrated directly in power dissipated in this load.

and the rectifier instrument may also be provided with a variety of multipliers of the type described in Sec. 4, to increase the power range that can be covered.

For general testing purposes it is desirable that the output meter be capable of supplying a wide range of load impedances, and of measuring a wide range of power. This can be accomplished with the circuit arrangement shown in Fig. 25. Here the impedance which the output meter offers to the source of energy is controlled by varying the taps on a transformer, while the power range can be varied by means of the attenuator placed between the rectifier instrument and the transformer secondary. In order to compensate for the variation of transformer losses at the different taps, series resistances are inserted so that a constant fraction of the total energy being measured will be delivered to the attenuator. The resistances in series with each primary tap are chosen so that the total direct-current resistance in the primary circuit is proportional to the square of the number of primary turns, while the resistances in series with the secondary taps are selected so that the sum of the secondary copper resistance, plus added resistance, is constant. The iron loss due to eddy currents (which is the principal part of the core loss) needs no compensation since it is a nearly constant fraction of the total loss at all times. The primary inductance must be large to extend the response to low frequencies, while the leakage reactance should be as small as possible to give a good high-frequency response. The resistance $R$ in series with the rectifier meter is to insure that the resistance across $aa$, looking to the right, will be constant in spite of variations in rectifier resistance with current; it should be five to ten times the rectifier resistance. This resistance $R_{aa}$ and the sensitivity of the D'Arsonval galvanometer used in the rectifier instrument control the maximum

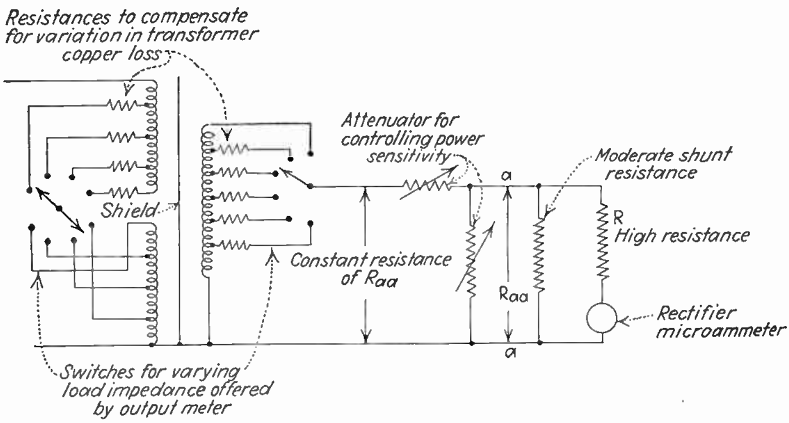


FIG. 25.—Circuit diagram of power-level meter capable of covering a wide power range, and of offering a variety of load impedances.

power sensitivity obtainable. The attenuator can be of the L type (see Sec. 23) and must be designed to have a constant input resistance of $R_{aa}$.

*Power-level Indicators.*—Circumstances often arise where it is necessary to monitor the power being carried by voice-frequency circuits to insure adequate volume while avoiding overloading. The simplest method of accomplishing this is to bridge a high-resistance rectifier voltmeter across the circuit as shown in Fig. 26*a*. By making the series resistance of the voltmeter high compared with the load resistance of the line, the shunting effect and consequent loss of power is negligible. At the same time the power sensitivity can be made high, if desired, by using a sensitive galvanometer $G$ in the output of the copper oxide rectifier unit. Where extreme sensitivity is essential, a transformer can be used to match the copper oxide rectifier unit to the line, as shown in Fig. 26*b*, thereby delivering to the rectifier all the power consumed in the bridging circuit rather than wasting most of it in a series resistance. It is generally desirable that monitors of voice-frequency currents give an indication of *average syllabic power*, while not indicating transient peaks of very short duration. This result is obtained by choosing a galvanometer $G$ in Figs. 26*a* and 26*b* that is rather heavily damped.

A vacuum-tube voltmeter arrangement such as illustrated in Fig. 26*c* is sometimes employed for monitoring volume level. The circuit elements $L$ and $C$, together with the damping of the meter $M$, determine the time interval over which the power is

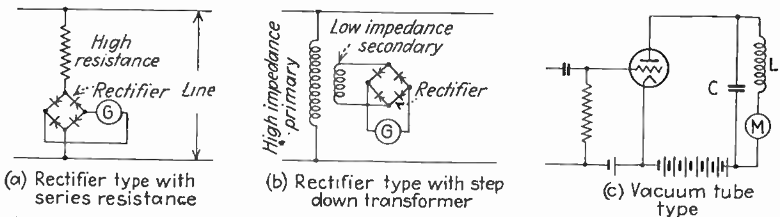



FIG. 26.—Various types of volume-level indicators for monitoring power level in voice-frequency circuits.

averaged to give an indication. Typical values of $L$ and $C$ are 300 henrys and 2 $\mu$f, respectively.

Instruments for monitoring power level are normally calibrated in decibels above or below some arbitrary power level. The standard level commonly employed is a power of 6 milliwatts, corresponding to 1.73 volts (effective) across a line having a 500-ohm load resistance (*i.e.*, a "500-ohm" line). It will be noted that the calibration depends upon the load resistance of the line across which the monitor is connected.

*Transient Peaks.*—Peak power in a circuit is measured by determining the crest voltage developed across a load of known resistance. Methods of determining transient voltage peaks are shown in Fig. 27. At Fig. 27*a* a condenser is charged to the peak voltage of the transient through a rectifier tube. In this way a charge once delivered to the condenser is trapped there and cannot escape except by leaking away. If the leakage resistance is high, the charge will remain long enough to be measured by a vacuum-tube voltmeter connected across the condenser. In

Fig. 27*b* a series of thyratron or similar gaseous discharge tubes are operated with a common grid-bias voltage but they have different plate voltages, so that they ignite at different applied voltages. Thus in Fig. 27*b* the first tube might be set to ignite at 10 volts, the second at 12, the third at 14, the fourth at 16, etc. These tubes can be ignited by a transient peak of extremely

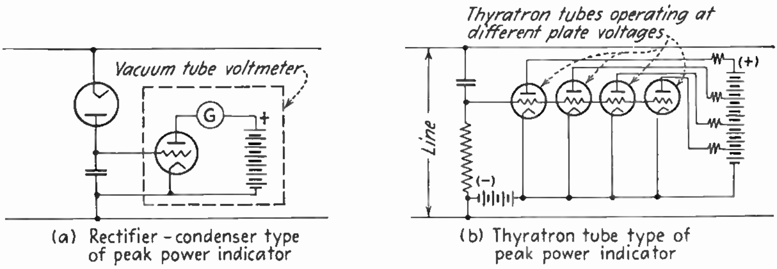



Fig. 27.—Types of peak-power indicators.

short duration and will then continue to glow until the plate voltage has been removed. Thus if a transient peak ignited the first three tubes and none of the remainder, one would know that the amplitude was at least 14 volts and less than 16. If it is merely necessary to give warning when the peak exceeds some predetermined value, a single thyratron set at the appropriate grid and plate potentials can be used.

## CHAPTER II

# CIRCUIT CONSTANTS AT LOW FREQUENCIES

**10. Direct-current Resistance.**—Resistance to direct current can be measured on a Wheatstone bridge, by the voltmeter-ammeter method, or with an ohmmeter. The Wheatstone bridge is the most accurate method and is standard for general laboratory use. The circuit diagram and constants of a typical Wheatstone bridge are shown in Fig. 28. The entire instrument, including galvanometer and battery, is normally contained in a single box. An unknown resistance is measured by connecting it at $R_x$ of the bridge and proceeding as follows: The ratio $R_a/R_b$ is first set so that the largest possible value of $R_s$ will be required for balance, after which $R_s$ is varied until balance is obtained. If the approximate value of the resistance $R_x$ is unknown, the proper ratio $R_a/R_b$ must be determined by trial and error. In obtaining a particular ratio of $R_a/R_b$ it is important that the largest possible values of resistances be used. Thus while a ratio of 10 could be obtained by making $R_a/R_b$ equal to 10/1, 100/10 or 1000/100 in the bridge of Fig. 28, the last combination is preferred. Preliminary adjustments should be made with the galvanometer shunted, and the shunt removed only to complete the balance. An answer containing four significant figures can normally be expected. When the resistance of circuits containing inductance or capacity is measured, it is necessary to press the battery key before pressing the galvanometer key in order that transients will not give the galvanometer needle an initial spurious kick.

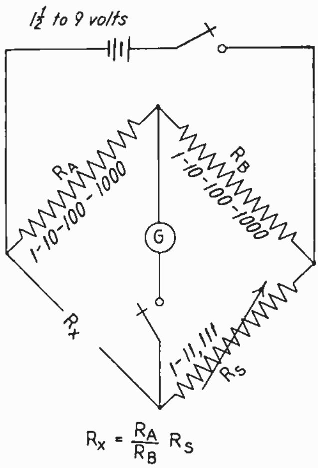



FIG. 28.—Circuit diagram and constants of typical Wheatstone bridge.

Special difficulties are encountered in bridge measurements of very low and very high resistances. With low resistances uncertainty is introduced by the resistance of leads and contacts. These troubles can be eliminated by using the Kelvin double bridge, a description of which can be found in any handbook. Accurate measurements of low resistances always require large currents in order that appreciable voltages may be developed across the bridge arms.

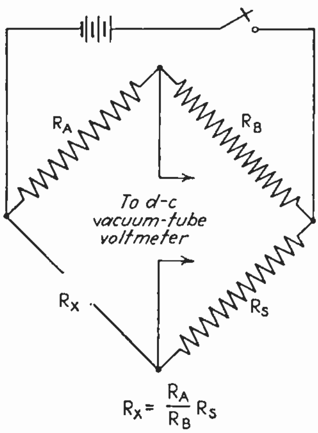



FIG. 29.—Circuit diagram of bridge suitable for the accurate measurement of high resistances from 1 to over 1000 megohms.

In measuring high resistance it is generally found that the sensitivity of the bridge galvanometer is not sufficient to permit accurate measurement. This is a result of the high ratio $R_a/R_b$ required in determining high resistances, and of the fact that no galvanometer has sufficient resistance to match the bridge impedance properly. These difficulties can be overcome by rearranging the bridge as shown in Fig. 29, so that the ratio $R_a/R_b$ is continuously variable, and by employing a fixed resistance of 1 megohm or more for the standard $R_s$. Balance is detected by means of a direct-current vacuum-tube voltmeter, using the simple circuit of Fig. 21*a* when the direct-current potential applied to the bridge is large (50 volts or more), or the highly stable voltmeter circuit of Fig. 21*b* with a very sensitive galvanometer when the applied voltage is small.

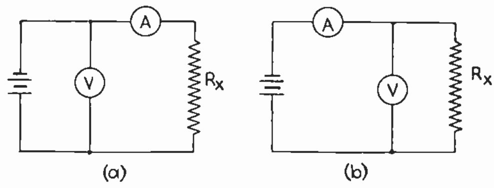



FIG. 30.—Circuit arrangements for measuring resistance by the voltmeter-ammeter method.

In the voltmeter-ammeter method of measuring resistance the voltage applied to the circuit and the resulting current are both measured with ordinary direct-current instruments. The ratio of voltage to current is then the desired resistance. In order to minimize errors resulting from the power consumed by the measuring instruments, the arrangement shown in Fig. 30*a* should be employed when the current drawn by the voltmeter is not negligible as compared with the current in the unknown resistance, while the circuit of Fig. 30*b* is required where the voltage drop in the ammeter *A* is not negligible compared with the applied voltage.

*Ohmmeters.*—Ohmmeters are suitable for making approximate resistance measurements and are used in servicing communication equipment. The two circuit arrangements shown in Fig. 31

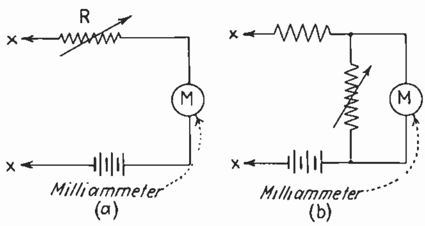



FIG. 31.—Ohmmeters used in servicing communication equipment.

can be used. The instrument of Fig. 31*a* operates on the assumption that the battery generates a constant voltage during its life but that the internal resistance of the battery increases with age. To operate, an initial adjustment is made by short-circuiting the terminals *XX* and adjusting the resistance *R* until full-scale reading is obtained on the milliammeter. When a resistance is now inserted between *XX*, the reading will be less than full scale by an amount that depends upon the resistance between *XX*, and the scale can be calibrated directly in ohms. Different ranges can be provided by various combinations of resistance *R*, number of cells in the battery, and shunts across the instrument.

The ohmmeter of Fig. 31*b* operates on the theory that, as the battery ages, its voltage drops but the internal resistance stays the same. In this instrument the initial adjustment is made by short-circuiting the prongs *XX* and varying the shunt of the milliammeter until full-scale deflection is obtained. Any resistance inserted in series with *XX* then causes the deflection to be less than full scale by an amount that depends upon the resistance, so that the instrument scale can be calibrated directly in ohms.

Ohmmeters at best are capable of only approximate measurements, since, as the battery ages, it changes both its internal resistance and generated voltage. A combination of the two methods shown in Fig. 31 would give the greatest accuracy.

**11. Alternating-current Bridges.**—The most satisfactory method of measuring resistance, capacity, inductance, and mutual inductance of a circuit at audio frequencies is by means of an alternating-current bridge. The schematic circuit diagram of such a bridge is shown in Fig. 32 and is similar to the direct-current Wheatstone bridge, except that the power source is now an oscillator instead of a battery and the galvanometer for detecting balance is replaced by telephone receivers. At balance the relation $Z_a/Z_b = Z_c/Z_d$ must hold, where the impedances $Z_a$, $Z_b$, $Z_c$, and $Z_d$ are now vector quantities.

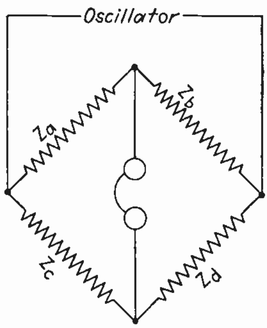



FIG. 32.—Schematic diagram of alternating-current bridge.

The four arms of an alternating-current bridge may be built up in a wide variety of combinations, as is apparent from Fig. 38 and Sec. 14. A large proportion of all bridge measurements can, however, be made with the arrangement shown in Fig. 33*a*. Here the arms $R_a$ and $R_b$ are resistances, $Z_x$ is the unknown, and $Z_s$ is a standard impedance. In this arrangement the unknown is measured against an impedance of the same kind; *i.e.*, a capacity is determined in terms of a standard capacity, resistance in terms of a resistance standard, etc. The equations for balance are

$$\left.\begin{aligned} Z_x &= \frac{R_a}{R_b} Z_s \\ X_x &= \frac{R_a}{R_b} X_s \\ R_x &= \frac{R_a}{R_b} R_s \end{aligned}\right\} \qquad (4)$$

where $Z_x$ = impedence being measured
$Z_s$ = impedance of standard
$R_x$ = resistance component of $Z_x$

$R_s$ = resistance component of $Z_s$
$X_x$ = reactance component of $Z_x$
$X_s$ = reactance component of $Z_s$.

It will be noted that the ratio between the unknown and standard impedances is determined by the resistance arms $R_a$ and $R_b$, which are often called ratio arms. The unknown and standard impedances must have identical power factors to give balance, and provision must accordingly be made in the bridge for equalizing the power factors.

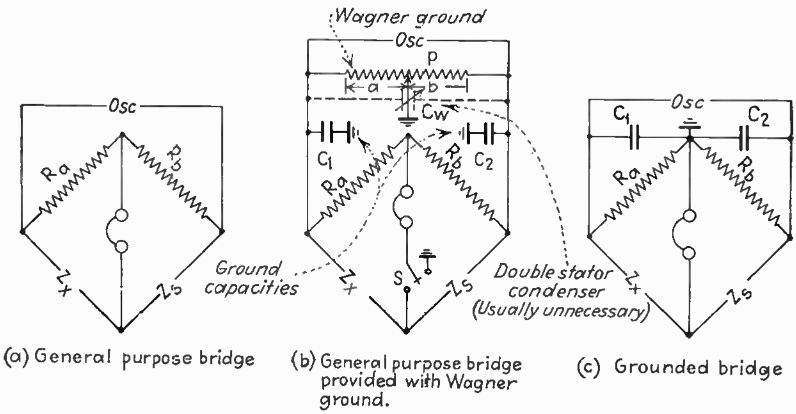



FIG. 33.—General-purpose form of alternating-current bridge, with and without Wagner ground.

The standard impedance $Z_s$ in Fig. 33*a* may be either fixed or variable. With fixed standards it is necessary to vary the ratio $R_a/R_b$ to obtain balance, while with a variable standard the ratio arms are set at appropriate values and then left fixed while the standard is adjusted.

*Wagner Ground Connection.*—When a bridge such as shown in Fig. 33*a* is used to measure high impedances (such as 50,000 ohms and up), the results will usually be in error and the balance point will be affected by placing the hands upon any part of the telephone receivers or their leads. This is caused by the fact that the neutral arm of the bridge is not at ground potential, and as a result spurious currents flow from the neutral arm to ground. Difficulties from this action can be eliminated by using a Wagner ground to bring the neutral arm of the bridge to ground potential, as shown in Fig. 33*b*. This consists of a potentiometer $P$ of perhaps 500 to 1000 ohms resistance. The principal ground capacities that cause trouble can be lumped together as $C_1$ and $C_2$ in Fig. 33*b*. The low-resistance sections $a$ and $b$ of the potentiometer $P$ are in shunt with these and so practically short-circuit them, causing the grounding point to be controlled by the slider on $P$ rather than the capacities to ground. By adjusting the slider so that $a/b = R_a/R_b$, the neutral arm is brought to ground potential, the bridge can be balanced without body effects, and the results will be correct.

When a bridge is equipped with a Wagner earth, the measuring procedure is as follows: The bridge is first balanced as well as possible without regard to the Wagner ground adjustment. The switch $S$ in Fig. 33*b* is then thrown so that the telephone receivers are connected between the neutral arm and ground, and the slider on the potentiometer adjusted until no sound is heard, which is an indication that the neutral arm is at ground potential. The switch $S$ is now thrown back to the original position, placing the receivers across the neutral arm, and the balance is completed. In rare cases it will be necessary to repeat this sequence of operations.

When a Wagner ground is applied to a bridge operating at radio frequencies, the reactance of capacities $C_1$ and $C_2$ is too low to be effectively short-circuited by the resistances $a$ and $b$ of $P$ (see Fig. 33*b*), and a double stator condenser $C_w$ must be included in the Wagner ground. This condenser is for the purpose of adding capacity in shunt with $C_1$ and $C_2$ until the total effective capacitive reactances to ground from the two sides of the bridge have the ratio $R_a/R_b = a/b$.

If an attempt is made to operate the bridge with one of the neutral corners grounded, as shown in Fig. 33*c*, the effect is to place the capacities $C_1$ and $C_2$ in shunt with the bridge arms. Unless these capacities are so proportioned that $C_2/C_1 = R_a/R_b$, the bridge will be unbalanced and errors introduced. In bridges having *fixed* ratio arms it is apparent then that by adjusting the capacities to the proper ratio a Wagner ground is not necessary.

*The Bridge Oscillator.*—Alternating-current bridges normally require perhaps 50 to 200 milliwatts for satisfactory operation in the range between 500 and 2000 cycles, when using a telephone receiver to detect balance. At higher and lower frequencies, either greater oscillator power or amplification of the bridge output is necessary to obtain reasonable accuracy. This is because the ear is less sensitive to low and high frequencies.

The extent to which the oscillator must deliver a pure sine wave depends upon whether or not the bridge balance is independent of frequency. When the balance is completely independent of the frequency, the bridge is balanced for the harmonics when it is balanced for the fundamental, and purity of wave form is unimportant. On the other hand, when the balance depends on frequency, the harmonics will still be heard when the proper balance point has been obtained, and may interfere greatly in carrying out the balance for the fundamental. This is particularly true when the fundamental is a low frequency, since then the ear is more sensitive to the harmonics than to the fundamental. Under such conditions it is highly desirable that the oscillator deliver a pure sine wave, or that a suitable low-pass filter be placed between the bridge output and the telephone receivers to cut out the harmonics.

*Bridge Calculations.*—It is sometimes necessary to determine the impedance which the null indicating device should have for maximum sensitivity, or again to know the current that will flow through the null device when there is a given degree of unbalance. Answers to problems of this sort can be most conveniently obtained with the aid of Thévenin's theorem. According to this theorem, *any electrical network no matter how complicated, having two output terminals, can, as far as current flowing in a load connected between these output terminals is concerned, be replaced by an equivalent generator, having a voltage V equal to the open-circuit voltage at the output terminals, and an internal impedance equal to the impedance looking into the network from the output terminals with all sources of voltage in the network short-circuited.*[1]

When applied to determining the current in the neutral arm of a bridge, this leads to the equivalent circuit of Fig. 34. The equivalent voltage $V$ is the voltage appearing across points $MN$ when the neutral arm is open, and it can be readily calculated. The equivalent impedance $Z_{eq}$ is the impedance across terminals

[1] For an elementary proof see page 55 of T. E. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Company, 1929. The only important limitation to this theorem is that the network be linear, *i.e.*, that current be everywhere proportional to voltage.

*MN* when the voltage source *E* is short-circuited,[1] and it can likewise be easily calculated. After the bridge is reduced to the simple equivalent circuit, the current through the neutral arm can be readily calculated. In order that the power delivered to the null indicator may be a maximum for a given degree of unbalance, it will be noted that the impedance of the indicating device should equal in magnitude the equivalent impedance $Z_{eq}$.

**12. General-purpose Alternating-current Bridge and Its Use to Measure Resistance, Capacity, and Inductance.**—The circuit details of a typical general-purpose alternating-current bridge are

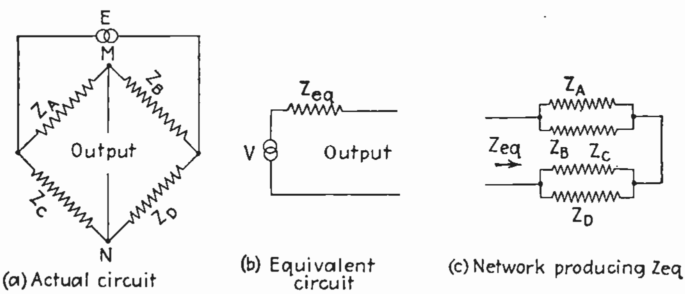



Fig. 34.—Application of Thévenin's theorem to simplify the bridge network.

shown in Fig. 35. The ratio arms are composed of a four-decade resistance $R_a$ and a resistance $R_b$ tapped at 1, 10, 100, 1000, and 10,000 ohms, thus permitting the use of a fixed or variable ratio. The third resistance $R_c$ is used as a standard in making resistance measurements and serves to equalize power factors in inductance and capacity measurements. The Wagner ground shown in Fig. 35 is usually mounted external to the bridge cabinet but is included in the diagram to show how the single switch *S* can be used to control the Wagner ground and to place the resistance $R_c$ in series with either the "standard" or the "unknown" arms. If the bridge is shielded according to the principles outlined in Sec. 22, it can be used at the lower radio frequencies.

The resistances used in alternating-current bridges must have a minimum of inductive and capacitive effects associated with

[1] In order to simplify the problem, it is assumed that the internal impedance of the voltage source *E* is negligible. If the bridge is only moderately unbalanced, this assumption introduces negligible error in the value of $Z_{eq}$ even when the generator has a high internal impedance.

them, and the wire should have zero temperature coefficient. Such resistances are discussed in detail in Sec. 23.

*Procedure for Measuring Resistance.*—Resistance is measured by using the $R_c$ resistance as the standard. This is done by short-circuiting the terminals marked "standard" in Fig. 35 and throwing switch $S$ to point 1 (see Fig. 36*a*). The ratio $R_a/R_b$ is then set at the value which it is estimated will be required, and $R_c$ adjusted until balance is obtained. The unknown resistance is then

$$R_x = \frac{R_a}{R_b}R_c \qquad (5)$$

If the unknown resistance is in the order of 50,000 ohms or more, the use of the Wagner ground is desirable. Under most conditions a balance accurate to four significant figures should be obtained. That is, a one-step variation in the last $R_c$ decade should upset the balance. The ratio arms should always be chosen in such a way that this number of significant figures is obtainable.

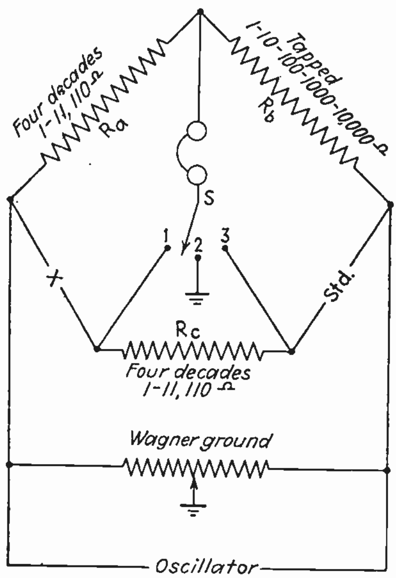



Fig. 35.—Circuit details of a typical general-purpose type of bridge equipped with Wagner ground.

When very high resistances are measured, it may be desirable to employ a fixed resistance standard of 100,000 or 1,000,000 ohms, in which case $R_c$ is set at zero. The balance is then obtained by varying the ratio $R_a/R_b$.

A bridge such as shown in Fig. 35 will measure resistances from a fraction of an ohm to over 1 megohm. The lower limit is set by errors and uncertainties caused by lead and contact resistance, coupled with lack of sensitivity of balance. The upper limit arises as a result of lack of sensitivity and because of complications introduced by stray capacities.

In measuring high resistances it will often be found that a perfect balance cannot be obtained even when a Wagner ground is used. This is because of slight inductive or capacity effects associated with the resistance, or stray capacity within the bridge itself. If one is interested only in the magnitude of the resistance and does not care to measure these reactive effects, it is permissible to obtain a null balance by shunting a variable condenser across either the unknown or standard resistance, as the situation requires.

The general-purpose bridge can be used as a direct-current Wheatstone bridge by substituting a galvanometer for the telephone receivers and a battery for the oscillator.

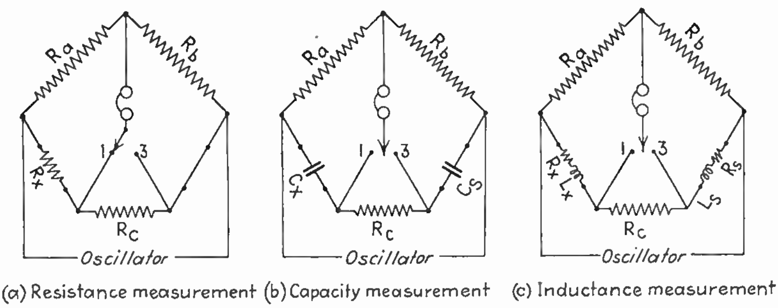



FIG. 36.—Circuit arrangements for measuring resistance, capacity, and inductance with the general-purpose bridge. A Wagner ground can be added when needed.

*Procedure for Measuring Capacity.*—To measure capacity the bridge is arranged as shown in Fig. 36*b*. The standard condenser $C_s$ is usually a fixed capacity so that balance is obtained by varying the ratio $R_a/R_b$, and using $R_c$ to equalize the power factors of the unknown and standard condensers. The balancing procedure is as follows: Resistance $R_c$ is set at zero, $R_b$ is set at what appears to be a suitable value, and the bridge is balanced as well as possible by varying $R_a$. The power factors are now equalized by placing the resistance $R_c$ in series with the condenser having the lowest losses, and varying $R_c$ until the best balance is obtained. One then readjusts first $R_a$, and then $R_c$, until a perfect null balance results. The proper location of $R_c$ is determined by throwing switch $S$ first to point 1, and then point 3. At balance the following formulas hold:

$$C_x = \frac{R_b}{R_a} C_s \tag{6a}$$

$$R_x = \frac{R_a}{R_b}R_s + R_c \tag{6b}$$

or

$$R_x = \frac{R_a}{R_b}R_s - R_c \tag{6c}$$

where $R_x$ and $R_s$ are the resistances of the "unknown" and "standard" condensers, respectively. Equation (6*b*) is used when the resistance $R_c$ is in series with the standard condenser while Eq. (6*c*) is used when $R_c$ is in series with the "unknown" condenser.

When a variable standard condenser is employed, the procedure is much the same except that the ratio arms are set at a fixed value and the standard instead of $R_a$ is varied to obtain balance.

Capacities from about 0.001 $\mu$f upward can be accurately measured by the procedure outlined above, but attempts to measure smaller capacities will usually lead to erroneous results because the stray capacities of the bridge arms and wiring are not negligible in comparison. In order to determine capacities smaller than about 0.001 accurately when using the bridge, one must employ the substitution method of measurement which is described in detail in Secs. 16 and 18.

*Procedure for Measuring Inductance.*—The circuit arrangement for inductance measurements is shown in Fig. 36*c* and differs from that used in determining capacity only in that an inductance standard is used. This standard may be either a fixed or variable inductance, although, as explained below, a variable standard is greatly to be preferred. The balancing procedure is similar in all respects to that used in the corresponding case in measuring capacity. About the only difference arises from the fact that inductances usually have much higher power factors (lower $Q$) than condensers, so that more attention must be paid to equalizing the power factors. When balance is obtained, the following formulas hold:

$$L_x = \frac{R_a}{R_b}L_s \tag{7a}$$

$$R_x = \frac{R_a}{R_b}(R_s + R_c) \tag{7b}$$

or

$$R_x = \frac{R_a}{R_b}R_s - R_c \tag{7c}$$

where

$L_x$ = unknown inductance
$L_s$ = standard inductance
$R_x$ = resistance of unknown inductance
$R_s$ = resistance of standard inductance
$R_a$, $R_b$, and $R_c$ = resistance of bridge arms.

Equation (7*b*) is used when balance is obtained with $R_c$ in the standard arm, while Eq. (7*c*) applies when $R_c$ is in the unknown arm.

A variable standard is very desirable in inductance measurements because it makes the balance for inductive and resistance components independent of each other. That is to say, when a variable standard is employed, the adjustment of the standard which gives minimum sound when the resistance balance is not correct, will be found to be the correct adjustment when the resistance balance has been completed, and vice versa.

With a fixed standard, it is necessary to vary the ratio of the bridge to obtain balance. The procedure for balancing is therefore to set the ratio at what is estimated to be the proper value and adjust $R_c$ as well as possible. It will normally be found that a perfect balance is not obtained so a new value of ratio must be tried and $R_c$ again adjusted, and so on over and over again. This procedure is slow because each time the ratio is changed $R_c$ must be readjusted. It requires considerable skill to adjust a bridge under these circumstances, and an inexperienced person will commonly be satisfied with what appears to be a slightly imperfect balance, but which actually represents considerable error. This difficulty is not encountered in making measurements of capacity using fixed standards, since here the resistance components are usually so small as to have almost negligible effect upon the balance.

The accurate measurement of small inductances, such as those employed as tuning inductances in broadcast and short-wave radio receivers, presents special problems. This is because the unknown inductance is so small that the inductance of the bridge wiring and resistance arms is not negligible in comparison, and because at audio frequencies the inductive reactance of the coil is usually less than the coil resistance. The best accuracy under such conditions is obtained with the substitution method of measurement using a bridge having inductance-compensated resistances.[1] Approximate results can be obtained by employing a variable standard and following the usual procedure for measuring inductances.

Difficulties are also encountered when the attempt is made to measure large inductances with the general-purpose bridge. In the first place, standards of high inductance are usually not available, which often makes it more convenient to use other types of bridges, such as the Hay bridge described in Secs. 14 and 15. In the second place, the distributed capacity of a large inductance is often sufficient to make the apparent inductance as measured differ appreciably from the true inductance. The self-capacity of a coil is in shunt with the inductance and thus makes the coil equivalent to a resonant circuit, with the result that the apparent inductance as measured at the coil terminals is larger than the true inductance unless the frequency is far removed from the frequency at which the coil is in parallel resonance. It can be shown that at frequencies that do not exceed 80 per cent of this frequency of self-resonance[2]

$$\left.\begin{array}{l}\text{Apparent inductance of coil}\\ \text{with self-capacity}\end{array}\right\} = \frac{L}{1 - m^2} \qquad (8a)$$

$$\left.\begin{array}{l}\text{Apparent resistance of coil}\\ \text{with self-capacity}\end{array}\right\} = \frac{R}{(1 - m^2)^2} \qquad (8b)$$

where $L$ and $R$ are the true inductance and resistance, respectively, and $m$ is the ratio of actual frequency to the frequency at which the self-capacity is in resonance with the coil inductance. Substitution in Eq. (8) shows that, if the error in inductance is to be kept below 1 per cent, measurements must be made at less than one-tenth of the natural resonant frequency of the coil ($m = 0.1$), or corrections calculated by Eq. (8) must be applied.

**13. Mutual Inductance and Coefficient of Coupling.**—The usual procedure for measuring the mutual inductance between two coils consists in connecting the two coils in series and measuring

[1] In such bridges, arrangements are provided for switching a low-resistance inductance into the circuit every time a resistance is cut out, and vice versa. In this way it is possible to keep the total inductance constant while varying the resistance. Inductance-compensated slide-wire potentiometers can also be constructed by using the same principle. See R. F. Field, Constant-inductance Resistors, *General Radio Experimenter*, vol. 8, p. 6, March, 1934.

[2] For further discussion see p. 61 of F. E. Terman, "Radio Engineering," McGraw-Hill Book Company, Inc., 1932.

the total inductance of the combination, after which the terminals of one coil are reversed and the measurement repeated. The mutual inductance is then one-fourth of the difference of the two measured inductances.[1]

In the case of an auto transformer, such as illustrated in Fig. 37, the foregoing procedure is not possible. In this case one can determine mutual inductance by measuring the impedance at the terminals 12 when the secondary terminals 34 are open, and when they are short-circuited. This, coupled with a measurement of the impedance at 34 with the primary terminals 12 open, will give all of the data needed to calculate the mutual inductance. The formula, assuming negligible resistance as compared with the reactance, is then[2]

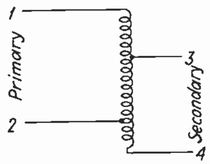


FIG. 37.—Auto-transformer coupling.

$$M = \sqrt{(L_p - L_c)L_s} \tag{9}$$

where $M$ is the mutual inductance between primary and secondary portions of the auto transformer, $L_c$ is the inductance at terminals 12 with the secondary terminals 34 short-circuited, $L_p$ is the inductance at terminals 12 with the secondary open (the primary inductance), and $L_s$ the inductance measured at 34 when the primary terminals 12 are open.

[1] This follows from the fact that, when the coils are in series aiding, the total inductance is $L_p + L_s + 2M$; while when the coils are in series opposition, the inductance is $L_p + L_s - 2M$. The difference between these values is then $4M$.

Other methods of measuring mutual inductance that can be used include the use of mutual inductance bridges (see Henney, "Radio Engineering Handbook" pp. 45–46, 1933), and the measurement of the voltage induced in the secondary by a known primary current. In this latter method the induced voltage is $\omega M I_p$ so $M$ is readily calculated when $I_p$, the frequency, and the induced voltage are known.

[2] Following the analysis given on p. 68 of the author's "Radio Engineering," the difference $L_p - L_c$ between primary inductance when the secondary is open and when it is short-circuited is the coupled inductance, which is $(\omega M)^2/\omega^2 L_s$. That is, $L_p - L_c = M^2/L_s$, and Eq. (9) follows at once. If the resistances are not negligible compared with the reactances, then the situation becomes rather involved because there is resistive as well as inductive coupling. The situation can be analyzed by the same type of approach as in other coupled circuits, using a term "mutual impedance" to replace the $(j\omega M)$ that appears in the usual case.

When the mutual inductance $M$ and the self-inductances $L_p$ and $L_s$ of the two coils are known, the coefficient of coupling can then be calculated by the formula

$$\text{Coefficient of coupling} = k = \frac{M}{\sqrt{L_p L_s}} \tag{10}$$

The coefficient of coupling represents the ratio of the actual mutual inductance to the maximum possible mutual inductance which can be obtained with the primary and secondary inductances.

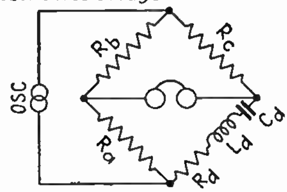
**14. Miscellaneous Types of Bridges.**—The general-purpose bridge that is described above, while capable of serving for most of the bridge measurements which must be made around a laboratory, represents only one of a large number of possible bridge circuits. Some of the more important of these bridge circuits are shown in Fig. 38, together with their equations for balance. Each of the bridge circuits shown in Fig. 38 has one or more limited fields of usefulness in which it is superior to most other types.

In the resonance bridge (Fig. 38*a*), balance is obtained by resonating the inductance and capacity to the applied frequency, thus making the $D$ arm equivalent to a resistance which can be balanced by $R_c$. This bridge is a special form of the general-purpose bridge and a Wagner ground may be used. The resonance bridge can be used to measure frequency in the audio-frequency range in terms of inductance and capacity (see Sec. 34). It is also sometimes employed to measure impedances having a capacitive reactance.[1]

The Hay bridge (Fig. 38*b*) compares an inductance with a capacity and finds its principal use in the measurement of incremental inductance (see Sec. 15 in which this bridge is considered at greater length). The balance of this bridge is dependent on the frequency, so that the bridge can be used to measure frequency (see Sec. 34). The Owen bridge (Fig. 38*c*) also measures inductance in terms of a capacity, but in this

[1] The Western Electric 1-B impedance bridge, which is widely distributed among universities and is used for transmission-line impedance measurements, operates as a resonance bridge when measuring capacitive reactances. When measuring inductive reactances, it is rearranged to be a modified form of the general-purpose inductance bridge.

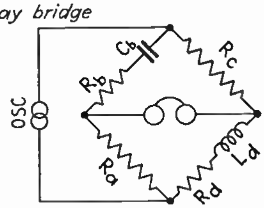
(a) *Resonance bridge*




$$wL = \frac{1}{wC}$$

$$R_d = \frac{R_a}{R_b} R_c$$

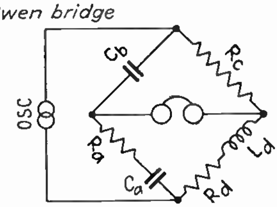
(b) *Hay bridge*




$$L_d = \frac{R_a R_c C_b}{1+(R_b w C_b)^2}$$

$$R_d = \frac{R_a R_b R_c (w C_b)^2}{1+(R_b w C_b)^2}$$

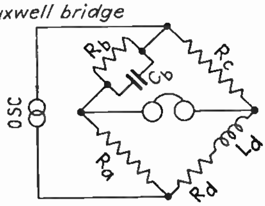
(c) *Owen bridge*




$$L_d = R_a R_c C_b$$

$$R_d = \frac{C_b}{C_a} R_c$$

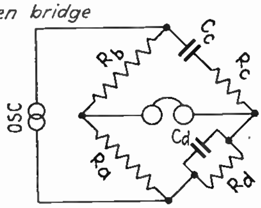
(d) *Maxwell bridge*




$$L_d = R_a R_c C_b$$

$$R_d = \frac{R_a}{R_b} R_c$$

(e) *Wien bridge*

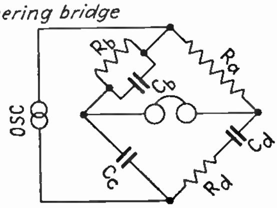



$$w^2 = \frac{1}{R_d R_c C_d C_c} \text{ and } \frac{C_d}{C_c} = \frac{R_b}{R_a} - \frac{R_c}{R_d}$$

or

$$C_d^2 = \frac{R_b R_d - R_a R_c}{R_a R_d^2 R_c w^2} \text{ and } C_c^2 = \frac{R_a}{(R_b R_d - R_a R_c) R_c w^2}$$

(f) *Schering bridge*




$$C_d = \frac{R_b}{R_a} C_c$$

$$R_d = \frac{C_b}{C_c} R_a$$

FIG. 38.—Miscellaneous types of bridge circuits.

bridge the arms are so arranged that the balance equations for both resistance and inductance are independent of frequency provided losses in condenser $C_b$ are negligible. The resistance and inductance balances may be made independent of each other by varying $R_a$ to obtain the inductance balance, and either varying $C_a$ or adding resistance to $R_d$ to obtain the resistance balance. The Maxwell bridge shown in Fig. 38*d* likewise compares an inductance with a capacity. The balance equations of this bridge do not involve the frequency, but the resistance and reactance balances are not independent unless $C_b$ is continuously variable.

Wien's bridge (Fig. 38*e*) measures capacity in terms of resistance and frequency. This bridge is useful because the standards of frequency and resistance are known to very great accuracy, thus making it possible to use this bridge for the precision determination of capacity.[1] The Wien bridge is also particularly suitable for measuring frequency in the audio-frequency range (see Sec. 34). The Wien bridge is a special form of the general-purpose bridge of Sec. 12 and can be used with a Wagner ground.

The Schering bridge shown in Fig. 38*f* is of importance because the ordinary capacity bridge when used to measure direct capacity becomes a Schering bridge (see Sec. 16).

**15. Measurement of Incremental Inductance.**—Incremental inductance is the inductance which is offered to the flow of an alternating current superimposed upon a direct current and is of fundamental importance in audio-frequency transformers and filter reactors. The most satisfactory method of measuring incremental inductance is with a Hay bridge arranged as shown in Fig. 39. With such a bridge, inductance is measured in terms of resistance and capacity, and balance is obtained by varying resistances $R_a$ and $R_b$. The direct current is introduced, measured, and controlled, in the neutral arm of the bridge, and is not affected by the process of obtaining a balance.

General-purpose bridges of the type described in Sec. 12 are not particularly well suited to incremental-inductance measurements. These bridges as normally constructed for general service are not designed to carry large currents such as are

[1] See J. G. Ferguson and B. W. Bartlett, The Measurement of Capacitance in Terms of Resistance and Frequency, *Bell System Tech. Jour.*, vol. 7, p. 420, July, 1928.

required to produce the necessary direct-current magnetization. Furthermore the direct current must be introduced in such a way that the process of balancing the bridge alters the amount of current that flows.

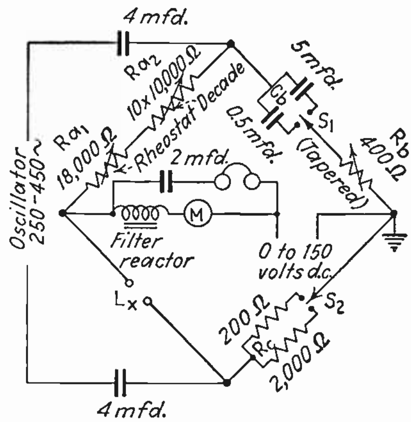



FIG. 39.—Hay bridge arranged to measure incremental inductance.

When the Hay bridge of Fig. 39 is adjusted to give balance, the following relations hold:

$$L_x = \frac{R_a R_c C_b}{1 + (R_b \omega C_b)^2} \tag{11a}$$
$$= \frac{R_a R_c C_b}{1 + \dfrac{1}{Q^2}}$$
$$R_x = \frac{R_a R_b R_c (\omega C_b)^2}{1 + (R_b \omega C_b)^2} \tag{11b}$$

where $Q$ is the ratio $(1/\omega C_b)/R_b = \omega L_x/R_x$.

When the losses in the inductance are reasonably low, as is nearly always the case, Eq. (11a) can be simplified to the following:

$$L_x = R_a R_c C_b \tag{12}$$

The extent of the approximation involved in Eq. (12) decreases as the $Q$ of the coil becomes larger. With $Q = 10$, Eq. (12) is 1 per cent high, while for $Q = 5$ the error becomes 4 per cent. It will be noted that to the extent that $1/Q$ is negligible the balance is independent of frequency.

In the practical determination of incremental inductance, one is interested primarily in the inductance and cares little if at all about the associated resistance. Furthermore the fact that the incremental inductance of a coil depends to an appreciable extent upon previous magnetic history and will change again with subsequent magnetic history means that great precision in incremental-inductance determinations is seldom required. Reproducible results can be obtained only by first demagnetizing the core and being careful to follow the same procedure in making each measurement. Because of this, Eq. (12) is entirely satisfactory for calculating the incremental inductance measured with the Hay bridge.

In making measurements of incremental inductance it is usually desirable to employ a relatively low frequency, such as 250 to 400 cycles, since at higher frequencies distributed capacity of the coil may cause the apparent inductance to differ appreciably from the true inductance (see Sec. 12).

When measurements on iron-core reactors using low frequencies are made, trouble will usually be encountered from harmonics generated either by the bridge oscillator or by the non-linearity of the magnetization curve of the core iron. Irrespective of their source, these harmonics tend to confuse the balance point since they are still audible in the telephone receivers when the bridge is properly balanced for the fundamental frequency. This makes it desirable to connect a low-pass filter between the bridge output and the telephone receivers to cut out these spurious frequencies.

*Constructional Features of Incremental-inductance Bridge for Laboratory Use.*—Although bridges such as shown in Fig. 39 for the measurement of incremental inductance are not at this writing commercially available, they can be readily assembled from standard parts. Circuit details including numerical values of circuit constants are shown in Fig. 39 for a bridge capable of measuring inductances up to 1000 henrys. In the bridge the inductance balance is obtained by varying $R_a$, which consists of a decade resistance unit having 10,000-ohm steps supplemented by a high-grade wire-wound rheostat to give continuous adjustment. A 4-in. metal dial, covered with white paper on which is a hand-calibrated scale marked in thousands of ohms, should be provided for the rheostat. The two condensers marked $C_b$

are for covering different ranges and can be built up to the correct capacity by combining commercial paper condensers. The rheostat $R_b$ is preferably tapered, can be any inexpensive wire-wound type, and need not be calibrated as it is used only to give the resistance balance. The resistances $R_c$ must carry the full direct current to be passed through the inductance being measured, and so must have a generous wattage rating. They are most satisfactorily made by padding out vitreous-enamel units which have resistances slightly below the required value. The blocking condenser and filter reactor in the neutral arm are for permitting alternating current to flow in the telephone receivers while blocking out the direct current. The meter $M$ for measuring the direct current should have an unshunted sensitivity of perhaps 10 ma, and shunts to extend the range as required. Any convenient source of direct current is suitable and should be provided with coarse and fine control so that a current of 1 or 2 ma through an audio transformer can be adjusted as accurately as 250 ma for a filter reactor. The blocking condensers in the oscillator circuit are to block direct current out of $R_a$ and the oscillator.

Grounding the bridge, as shown, places one terminal of the inductance at ground potential for direct current. If the condensers $C_b$ are mounted in metal cans, care should be taken to insulate the cases from ground, since otherwise the capacity between the condenser and its grounded container will place a capacity across $R_a$ that may be sufficient to affect the bridge behavior adversely.

The bridge as shown in Fig. 39 provides three ranges as follows:

| Range | Maximum inductance, henrys | $C_b$, microfarads | $R_c$, ohms |
|---|---|---|---|
| 1 | 11.8 | 0.5 | 200 |
| 2 | 118 | 5.0 | 200 |
| 3 | 1180 | 5.0 | 2000 |

It will be noted that for Range 2 the inductance in henrys is equal to $R_a$ in thousands of ohms, thus making the bridge direct reading if suitably calibrated. The necessary switching for

Ranges 1, 2, and 3 can be accomplished by a single three-position key controlling both $S_1$ and $S_2$ and giving multiplying factors of 0.1, 1.0, and 10, respectively.

The alternating flux density in the unknown inductance can be determined by measuring the voltage drop across it with a vacuum-tube voltmeter provided with leak and condenser to block off direct-current potentials from the grid. A knowledge of voltage drop, frequency, incremental inductance, and number of turns then makes possible a calculation of alternating flux.

*Incremental Inductance with High Alternating Flux Densities.* The bridge described in Fig. 39 is designed for making measurements when the alternating voltage across the unknown inductance does not exceed 50 to 100 volts. In order to determine incremental inductance with very high alternating flux densities such as exist in the input inductance of a choke-input filter, the bridge circuit of Fig. 39 must be redesigned so that alternating potentials of the order of 500 to 1000 volts may be applied without overheating or breaking down parts of the bridge. Excitation for the bridge is preferably 60 cycles, since this gives higher flux densities in proportion to voltage than higher frequencies, and balance can be detected with a vacuum-tube voltmeter. The applied voltage can be varied by means of a Variac[1] operating on the primary of a step-up transformer.

An alternative method that can be used to determine incremental inductance at high flux densities with fair accuracy is shown in Fig. 40. Here the inductance under test is actually used as the first choke of a choke-input filter system. The direct-current magnetization is controlled by varying the load resistance $R$. The incremental inductance is determined from measurements of the alternating voltage developed across half of the transformer secondary, and the alternating current which flows through the filter condenser $C$. Both of the measurements can be made with rectifier-type instruments. It can be readily shown that with a single-phase full-wave rectifier the effective value of the fundamental component of the ripple voltage that is applied to the input of the choke is 0.424 of the effective alternating voltage $E_{ac}$. Virtually this entire voltage is applied

[1] A Variac is an auto transformer in which the secondary voltage is adjusted by a high-resistance sliding contact. It was developed by the General Radio Company and is made in several sizes.

to the input inductance and nearly all of the resulting current flows through the first filter condenser. While there are other frequencies applied to the first choke, these are comparatively small and also produce proportionally less current because of their higher frequency. Hence to a fair degree of accuracy the inductance of the input choke is

$$\text{Incremental inductance} = \frac{0.424E_{ac}}{I_c} \tag{13}$$

where $E_{ac}$ is the r.m.s. voltage across half of the transformer secondary, while $I_c$ is the r.m.s. current flowing in the first filter condenser. The alternating flux density can be varied as desired by controlling the primary voltage of the rectifier transformer.

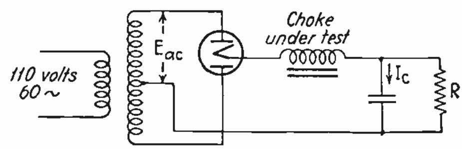



FIG. 40.—Circuit for approximate measurement of incremental inductance at high alternating flux densities.

**16. Special Problems Involved in the Measurement of Capacity.** *The Substitution Method.*—The substitution method is the most satisfactory method of measuring capacities of 0.001 $\mu$f and less. It can be applied to any bridge circuit capable of measuring capacity, and yields results that have an accuracy limited only by the accuracy of the calibration curve of a variable standard condenser.

The circuit arrangements for applying the substitution method to the general-purpose bridge of Sec. 12 are shown in Fig. 41. Here the capacity $C_x$ to be measured is connected in parallel with a variable standard condenser $C_s$, while the capacity $C_c$ is any available capacity a little larger than the unknown capacity ($C_c$ can be an ordinary variable condenser). The bridge is brought into balance by varying the standard capacity $C_s$, and equalizing the power factors by resistance $R_c$ until balance is obtained. The unknown capacity $C_x$ is then disconnected, and the capacity of $C_s$ increased until balance is restored. The increase in capacity required is obviously exactly equal to the capacity $C_x$ of the condenser being measured, *and the result is independent of any bridge errors, since the bridge conditions do* *not vary.* The accuracy hence depends only upon the calibration of the standard condenser $C_s$. The equivalent series resistance of the unknown capacity can be determined by using a standard condenser that has losses which are independent of the condenser setting. To do this, one sets an auxiliary resistance $R_s$ (shown dotted in Fig. 41) at zero for the preliminary balance, when both $C_s$ and $C_x$ are present, and then uses $R_s$ to compensate for the losses removed when $C_x$ is disconnected. Then

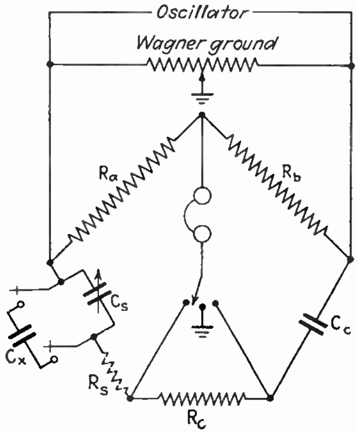



FIG. 41.—Capacity measurements with general-purpose bridge using substitution method.

$$\left.\begin{array}{l}\text{Equivalent series resistance}\\ \text{of unknown capacity } C_x\end{array}\right\} = \left(\frac{C_s}{C_x}\right)^2 R_s \quad (14)$$

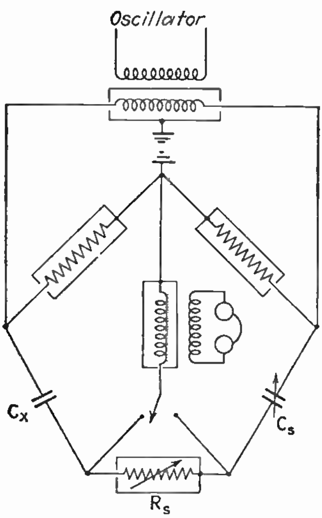



FIG. 42.—Shielded capacity bridge. The entire bridge is inclosed in a copper-lined box that serves as the ground.

where $C_s$ is the capacity of the standard *after* disconnecting the capacity $C_x$, and $R_s$ is the resistance that must be added in series with $C_s$ to compensate for the reduction in losses that results when $C_x$ is disconnected.

*Capacity Bridges.*—The accurate measurement of capacities smaller than 0.001 $\mu$f by direct measurement, as contrasted to the substitution method, requires modifications in the general-purpose bridge in order to avoid errors from stray bridge capacities. The essential features of such a modified bridge are shown in Fig. 42. This is a unity-ratio shielded bridge with the junction of the ratio arms grounded. In order that the ratio may be exactly unity, the two ratio arms are identical and are mounted symmetrically in grounded shields. The oscillator input to the bridge is supplied through a shielded transformer provided with a symmetrical secondary having equal capacities from each side to ground, so that the capacities in shunt with the ratio arms are identical and are constant. The unknown capacity $C_x$ is measured in terms of a variable standard capacity $C_s$. The shielded decade resistance $R_s$ is for the purpose of equalizing power factors. The telephone receivers are connected to the neutral arm through a shielded impedance-matching transformer. Care must be taken to prevent magnetic coupling between the input and output transformers, by using transformers inclosed in iron cases, and having a semi-astatic type of winding. Any residual coupling can be eliminated by proper orientation of the transformers. The entire bridge should be enclosed in a copper-lined box to prevent stray couplings to neighboring objects.

A capacity bridge such as described above finds its chief usefulness where so many capacity measurements are to be made that a special bridge devoted exclusively to this purpose is justified. Otherwise a general-purpose bridge employing the substitution method is preferable because of the greater usefulness of the bridge.

*Direct Capacity.*[1]—In communication work it is sometimes necessary to measure the capacity that exists directly between two electrodes when there is capacity between each of these electrodes and one or more other electrodes. Such a capacity is called *direct capacity.* A typical case is supplied by a vacuum tube, where each element of the tube has capacity to every other element, and also capacity to ground, as illustrated in Fig. 43.

Any one of the capacities illustrated in Fig. 43 can be measured independently of the other capacities by use of the capacity bridge of Fig. 42. The capacity to be measured is connected

[1] For further information on this subject, and particularly for additional ways of measuring direct capacity, see Harold A. Wheeler, Measurement of Vacuum-tube Capacities by a Transformer Balance, *Proc. I.R.E.*, vol. 16, p. 476, April, 1928; Lincoln Walsh, A Direct-capacity Bridge for Vacuum-tube Measurements, *Proc. I.R.E.*, vol. 16, p. 482, April, 1928; E. T. Hoch, A Bridge Method for the Measurement of Inter-electrode Admittance in Vacuum Tubes, *Proc. I.R.E.*, vol. 16, p. 487, April, 1928; Robert F. Field, Direct Capacity and Its Measurement, *General Radio Experimenter*, vol. 8, p. 5, November, 1933.

in the $X$ arm of the bridge and all remaining electrodes brought to the grounded junction of the ratio arms. This is shown in Fig. 44 for the measurement of $C_{gp}$. It will be noted that certain of the associated capacities are placed in shunt with the neutral arm where they have no effect, and the remainder placed in shunt with the ratio arm $R_a$. The result is the Schering bridge of Fig. 38*f*, and it will be noted from the resulting balance equations for Fig. 44 that the capacity in shunt with the $R_a$ arm affects only the resistance balance, which no longer can be used to determine the losses of the capacity being measured. As far as the capacity balance is concerned, the usual relation $C_x = (R_b/R_a)C_s$ still holds.

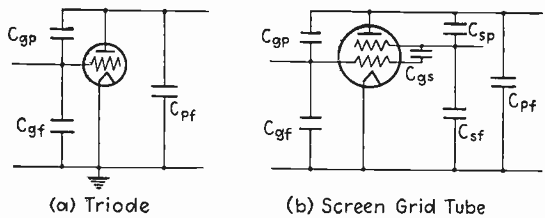



FIG. 43.—Direct capacities existing in triode and screen-grid tubes.

Direct capacity can also be determined by using a general-purpose bridge, provided the substitution method is employed. The equivalent series resistance of a direct capacity can be determined by using a Wagner ground, instead of grounding the bridge directly as shown in Fig. 44, and tying all of the unused electrodes to the Wagner ground. The Wagner ground in this case must be provided with balancing condensers as shown dotted in Fig. 33*b*.

In the event that one electrode of the direct capacity being determined is grounded, then the arrangement illustrated in Fig. 44 must be modified by grounding the end of the neutral arm opposite that shown grounded in Fig. 44. This permits one to connect the unknown capacity in the $X$ arm of the bridge. All other electrodes associated with the capacity under test are then gathered together and connected to the junction of the ratio arms. The procedure from here is then the same as in the case outlined above.

*Electrolytic Condensers.*—In measuring electrolytic condensers it is necessary to impress a direct-current polarizing voltage upon the condenser while making measurements, and it is generally desirable that means be provided for measuring the

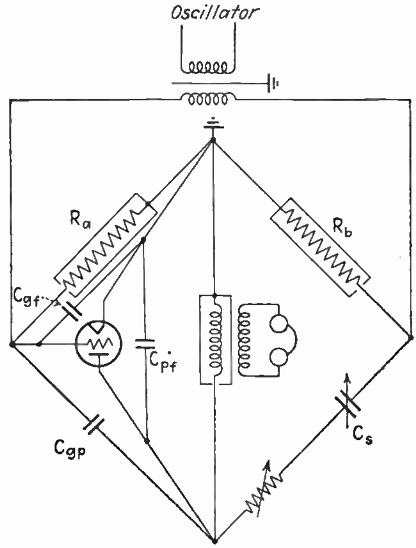



FIG. 44.—Schematic diagram showing the method of measuring direct capacity between two ungrounded electrodes using the capacity bridge of Fig. 42.

leakage current through the condenser. These features can be added to a bridge without difficulty, as illustrated in Fig. 45.

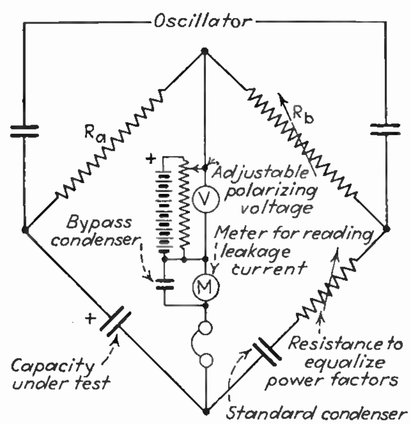



FIG. 45.—Arrangement by which an ordinary general-purpose bridge can be used in making measurements on electrolytic condensers.

**17. Capacity and Inductance Meters.**—Indicating-type capacity and inductance meters are sometimes employed in servicing radio equipment. These instruments utilize a 60-cycle source of voltage, in series with which is the unknown capacity (or inductance), a fixed resistance, and a rectifier-type microammeter, as shown in Fig. 46. The current that flows in such a circuit is obviously a function of the capacity (or inductance) in the circuit, so that the current instrument may be calibrated in terms of henrys or microfarads. The actual operating procedure is as follows: The leads to the inductance (or capacity) are short-circuited, and a shunt on the direct-current instrument in the rectifier output is adjusted until full-scale deflection is obtained. This takes care of variations in supply voltage. The short circuit across the unknown reactance is then removed, and the capacity (or inductance) is read on the instrument dial.

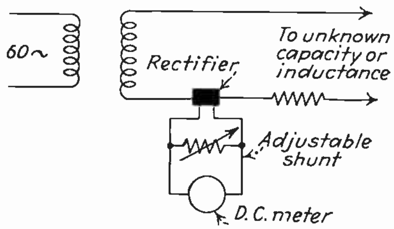



FIG. 46.—Simple type of capacity and inductance meter used in service work.

By using a potential of 100 volts with a microammeter having a sensitivity of 250 $\mu$a for full scale, capacities as small as 0.001 $\mu$f can be measured. Other ranges can be obtained by changing the supply voltage or the series resistance.

When used to measure inductances, instruments such as these are suitable for checking filter reactors, audio-transformer windings, etc., but cannot be used in determining the inductance of tuning coils, since these have negligible reactance at 60 cycles.

# CHAPTER III

## CIRCUIT CONSTANTS AT RADIO FREQUENCIES

**18. The Substitution Method of Measurement at Radio Frequencies.**—The simplicity and directness of the substitution method make this the most widely used of all measuring methods at radio frequencies. It is employed to measure the capacity of condensers, the inductance of coils, antenna constants, transmission-line impedances, etc. Although the exact details involved in its application vary according to the situation, an idea as to the possibilities of the method is brought out by considering a number of important examples.

*Measurement of Capacity by the Substitution Method.*—The capacity of variable condensers and small fixed condensers is often determined by the substitution method at radio frequencies, rather than by some type of audio-frequency measurements. The measuring procedure is essentially as follows: A calibrated variable condenser and the capacity to be measured are placed in parallel and used to tune a circuit to resonance at some convenient frequency. The unknown capacity is then disconnected, and the capacity of the calibrated condenser increased until the original resonant frequency is again obtained. The unknown capacity is now obviously equal to the change in capacity of the calibrated condenser. It will be noted that the accuracy of the result can be made very high as it depends only upon the accuracy of the calibration curve of the calibrated condenser, and the exactness with which the tuned circuit may be reset to the same resonant frequency.

The principal problem involved in the substitution method of measuring capacity is in determining when the circuit has been retuned to the resonant frequency that existed before the capacity being measured was removed. One method of accomplishing this is to couple a tuned circuit loosely to a low-power oscillator. As the tuned circuit is brought into resonance with the oscillator frequency, there is a jump in the d-c plate current drawn by the oscillator. By making use of this phenomenon to indicate when the tuned circuit and the oscillator are in resonance, and allowing the oscillator frequency to remain constant while making the measurements, it is possible to retune the resonant circuit with high accuracy. It will be noted that the oscillator frequency need only be constant, and that its actual value does not enter into the result.

A variation of this procedure is to utilize the circuit containing the calibrated and unknown capacity as the frequency-determining element of an oscillator. The oscillations thus generated are picked up as a squeal on an oscillating detector which is tuned to give zero beat. The calibrated condenser can then be reset with great accuracy to compensate for the removal of the unknown capacity by readjusting to regain zero beat while the oscillating detector is left untouched.

*Measurement of the Apparent Inductance of a Coil.*—The apparent inductance of a coil can be measured by connecting it across a calibrated variable condenser and determining the capacity required to make the resulting tuned circuit resonant at some particular frequency. According to the usual law of resonant circuits, one then has $\omega L = 1/\omega C$, or

$$L = \frac{1}{\omega^2 C} \tag{15}$$

The inductance measured in this way is the apparent inductance and may be somewhat higher than the true inductance because of distributed capacity, as explained in Sec. 12.

Another method of measuring the apparent inductance of a coil is as follows: A tuned circuit containing a calibrated variable condenser is first adjusted to resonance at the frequency desired. The coil to be measured is then connected in parallel with the tuned circuit (*i. e.*, across the condenser), after which the condenser is readjusted until the same resonant frequency as before is obtained. If the change in capacity of the standard condenser is $\Delta C_s$, then the reactance of the coil is $1/(\omega \Delta C_s)$ ohms, and the apparent inductance is $1/(\omega^2 \Delta C_s)$ henrys.

*Measurement of Radio-frequency Choke Coils.*—The impedance of a radio-frequency choke coil can be thought of as being produced by a reactance shunted by a resistance as shown in Fig. 47*a*, or a reactance in series with a resistance as in Fig. 47*b*. The equations relating these two methods of representing coil impedance are given in the figure.

The choke-coil reactance $X_1$ as illustrated in Fig. 47*a* can be obtained by the following procedure: A resonant circuit containing a calibrated variable condenser is tuned to the frequency at which the choke-coil characteristics are desired. The choke coil is then connected in parallel with the tuned circuit, *i.e.*, across the condenser, which is readjusted to restore the original resonant frequency. The equivalent reactance of the radio-frequency choke coil as represented in Fig. 47*a* is then

$$\text{Reactance } X_1 = \frac{1}{\omega \Delta C} \tag{16}$$

where $\Delta C$ is the change in capacity of the calibrated condenser required to compensate for the addition of the choke coil. The reactance is inductive when the capacity of the calibrated

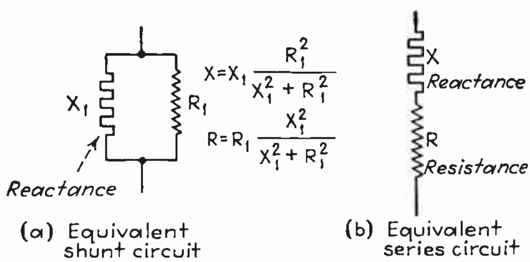



FIG. 47.—Methods of representing the impedance of a radio-frequency choke coil.

condenser must be increased after the coil is added, and it is capacitive if the capacity must be decreased.

The equivalent resistance $R_1$ of a radio-frequency choke coil as shown in Fig. 47*a* can be determined by measuring the parallel resonant impedance of the tuned circuit before and after the addition of the choke coil. The two measurements are, of course, made when the resonant frequency is the same. If the equivalent parallel resistances before and after the addition of the choke coil are $R'$ and $R''$, respectively, then[1]

[1] This follows from the fact that $R_1$ in parallel with $R'$ must give an equivalent resistance $R''$. That is,

$$\frac{R_1 R'}{R_1 + R'} = R''$$

This leads immediately to Eq. (17).

$$\text{Resistance } R_1 = \frac{R''R'}{R' - R''} \tag{17}$$

The parallel resonant resistances $R'$ and $R''$ can be measured directly by the resistance-neutralization method, described in Sec. 21, or can be calculated from a knowledge of the equivalent series resistance and the circuit $Q$.

The impedance of a radio-frequency choke coil is affected very greatly by small shunting capacities. As a result, care must be taken when the measurements are being made, and it must also be realized that the impedance which is measured is the impedance in that particular set-up and is not necessarily the same impedance that might be obtained with some other arrangement of leads, neighboring conductors, and ground.

*Antenna and Transmission-line Impedance.*—Measurements of antenna and transmission-line impedance by the substitution method are discussed in Secs. 65 and 67.

**19. Distributed Capacity of Coils.**—Every inductance has a small amount of capacity associated with it as a result of dielectric stress between various parts of the coil. This distributed capacity acts very much as though it were lumped across the terminals of the coil, and it is important because it limits the range over which a coil can be tuned with a given variable condenser. The distributed capacity also causes the coil to have an apparent inductance that differs from the true inductance as measured at low frequencies. This is caused by incipient parallel resonance between the coil inductance and its distributed capacity (see Sec. 12 for further discussion of this point).

The distributed capacity which may be considered as being lumped across the coil terminals can be determined by adding a known capacity in parallel with the coil and measuring the resonant frequency of the resulting tuned circuit. Knowing the true inductance of the coil and the frequency, one can calculate the total tuning capacity, from which by subtraction of the added capacity one obtains the distributed capacity. It is more accurate, however, to make tests with several added capacities. Then if one plots $1/f^2$ as a function of the added capacity, as shown in Fig. 48, a straight line will result. This line when extrapolated will intersect the capacity axis at a negative capacity which is equal to the distributed capacity, and the slope of the line is a measure of the true inductance of the coil according to the equation

$$\text{Coil inductance in henrys} = 0.0253m \qquad (18)$$

where $m$ is the slope of the curve of $1/f^2$ plotted against capacity, when $f$ is in megacycles and capacity is in micromicrofarads.[1]

The necessity of plotting a curve such as shown in Fig. 48 can be avoided by using a calibrated condenser to tune the coil to resonance with an oscillator, and then reducing the condenser capacity until the coil is brought to resonance with the

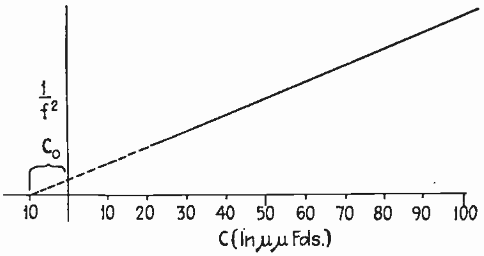



FIG. 48.—Plot of $1/f^2$, where $f$ is the resulting resonant frequency, as a function of external tuning capacity. The value of negative capacity $C_0$ at which the extrapolated line intercepts the capacity axis is the distributed capacity of the coil, and the slope of the line is a measure of the coil inductance.

second harmonic of the oscillator. If $C_1$ is the tuning capacity required for the fundamental frequency, and $C_2$ the capacity at the second harmonic, one then has[2]

[1] This is derived as follows: The relation between circuit constants and frequency in a resonant circuit is

$$2\pi fL = \frac{1}{2\pi f(C + C_0)}$$

where $C + C_0$ is the total tuning capacity (including distributed capacity $C_0$). Solving this equation for $1/f^2$ gives

$$\frac{1}{f^2} = 4\pi^2 LC + 4\pi^2 LC_0$$

This is the equation of a straight line intersecting the $C$-axis at $-C_0$, and having a slope of $(4\pi^2 L)$.

[2] Equation (19) is derived by the following reasoning: The actual tuning capacity effectively present in the two cases is $(C_1 + C_0)$ and $(C_2 + C_0)$, where $C_0$ is the distributed capacity. These capacities must be in the ratio of 4 to 1 since the tuning capacity is inversely proportional to the square of the frequency. That is,

$$(C_1 + C_0) = 4(C_2 + C_0)$$

$$\text{Distributed capacity} = \frac{C_1 - 4C_2}{3} \tag{19}$$

A method sometimes given for the determination of distributed capacity involves measuring the resonant frequency of the coil when tuned only by the distributed capacitance. This can be done by loosely coupling the coil to an oscillator and observing the frequency at which the coil reacts upon the oscillator to cause a sudden increase in d-c plate current. A knowledge of this resonant frequency and the true inductance of the coil will permit a determination of the apparent distributed capacity. The capacity obtained in this way will always be smaller than when the same capacity is measured by the preceding methods, because the voltage and current distribution in the coil with no external tuning condenser is quite different from the distribution in the presence of a normal tuning capacity. For this reason the distributed capacity of a coil should not be determined by the natural resonant-frequency method.

**20. Measurement of Very Small Capacities.**—The measurement of very small capacities and capacity changes can be most satisfactorily carried out by means such as illustrated in Fig. 49, where the desired capacity can be measured in terms of a much larger capacity. In Fig. 49 the unknown capacity is arranged so that it can be made part of the tuning capacity of an oscillator, and the size of the capacity is determined from the change in oscillator frequency which it produces. An auxiliary oscillator is used as a fixed standard, and the frequency variations of the measuring oscillator obtained by observing changes in the beat note as described in Sec. 35.[1] The relation between frequency change and capacity is given by the equation

$$C_x = C_0\left[\left(\frac{f_0}{f_1}\right)^2 - 1\right] \tag{20}$$

where $C_x$ is the unknown capacity, $C_0$ the original oscillator tuning capacity, $f_0$ the oscillator frequency when tuned only

and Eq. (19) follows at once by solving for $C_0$. This method is due to Batcher. See Ralph R. Batcher, Rapid Determination of Distributed Capacity of Coils, *Proc. I.R.E.*, vol. 9, p. 300, August, 1921.

[1] In noting the change in beat note, care must be taken to insure that the beat note does not go through zero as the frequency varies, or if it does to allow for this fact.

by $C_0$, and $f_1$ the oscillator frequency when tuned by $C_0$ and $C_x$ in parallel. In the special case where $C_x$ is small compared with $C_0$, Eq. (20) can be rewritten as

$$C_x = 2C_0\left(\frac{f_0 - f_1}{f_0}\right) \tag{20a}$$

The oscillator tuning capacity $C_0$ includes tube, wiring, and stray capacities and must be determined experimentally by using a known capacity for $C_x$ and noting the resulting frequency $f_1$. Since this calibrating capacity can be large enough to be

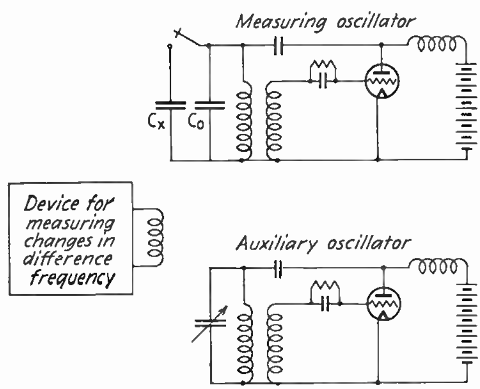



FIG. 49.—Evaluation of a small capacity in terms of the change in oscillator frequency which it produces.

measured very accurately by other methods, it is possible to evaluate $C_0$ very accurately.

This method is capable of measuring extremely small capacities. Thus if $C_0 = 50$ $\mu\mu$f and $f_0 = 10{,}000{,}000$, a capacity $C_x$ of $10^{-5}$ $\mu\mu$f will cause a change in frequency of one cycle per second, which is readily measured with accuracy if the frequencies $f_0$ and $f_1$ can be compared within a few seconds.

**21. Resistance of Tuned Circuits.**—The methods commonly employed to measure the resistance of tuned circuits are the resistance-neutralization, the resistance-variation, the capacity-variation, and frequency-variation methods. A number of other methods have been devised, but these are all less satisfactory for one reason or another and are seldom used.

*Resistance-neutralization Method of Measuring Tuned Circuit Resistance.*[1]—In this method, the negative resistance of a dynatron is connected in parallel with the tuned circuit being measured, as shown in Fig. 50. The magnitude of this negative resistance is varied by controlling the grid bias of the tube until oscillations are just on the verge of being started. With this condition existing, the negative resistance of the dynatron is

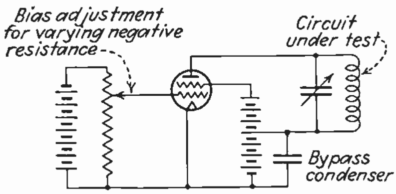



FIG. 50.—Schematic diagram illustrating the resistance-neutralization method of measuring the resistance of a tuned circuit.

exactly equal to the parallel-resonant impedance of the tuned circuit, so that

$$R_n = \frac{(\omega L)^2}{R_s} = \omega LQ$$

Hence

$$R_s = \frac{(\omega L)^2}{R_n} \tag{21a}$$

$$Q = \frac{\omega L}{R_s} = \frac{R_n}{\omega L} \tag{21b}$$

where $R_n$ is the negative resistance required to neutralize the resistance of the circuit, $R_s$ is the series resistance of the resonant circuit, $L$ is the true inductance, and $\omega$ is $2\pi$ times the resonant frequency. No error is caused by distributed capacity, stray wiring capacity, and tube capacity, as these merely assist the tuning condenser to bring the circuit to resonance. It is to be noted that the negative resistance $R_n$ of the dynatron can be measured at audio frequencies rather than at the frequency at which the circuit oscillates. The inductance of the circuit can likewise be measured at audio frequencies, preferably with the aid of a bridge, and as this is very far from the natural resonant

[1] See Hajine Iinuma, A Method of Measuring the Radio-frequency Resistance of an Oscillatory Circuit, *Proc. I.R.E.*, vol. 18, p. 537, March, 1930.

frequency formed by the inductance and its distributed capacity, no correction need be made for the latter.

The negative resistance can be supplied by any screen-grid tube which gives a reasonable amount of secondary emission at the plate.[1] The plate potential should approximate the

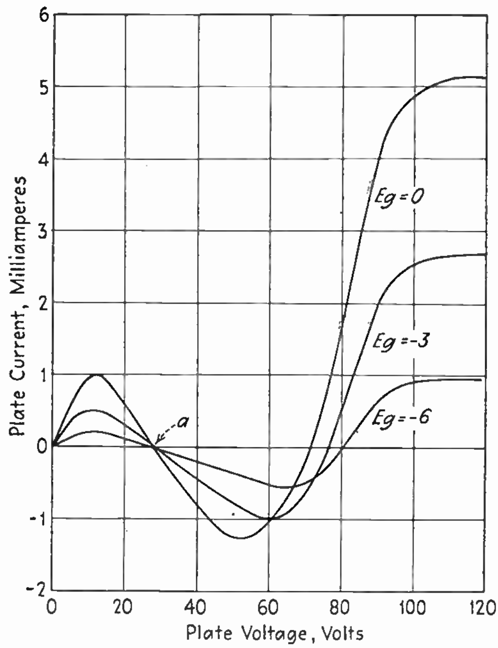



FIG. 51.—Dynatron characteristic of typical screen-grid tube.

value that places the operating point at the point of zero plate current indicated as *a* in Fig. 51. A screen-grid potential of 90 to 135 volts is recommended for ordinary tubes. To control the negative resistance the grid bias should be adjustable from approximately zero to cut-off bias.

*Incremental Method of Measuring Negative Resistance.*—The negative resistance developed by a dynatron may be measured

[1] The types 22 and 24 tubes are nearly always satisfactory dynatrons. Other screen-grid tubes, and pentodes with suppressor tied to accelerator grid, are satisfactory in some cases. Many of the tubes now on the market have had their plates so treated during exhaust as to give very little if any secondary emission, and so cannot be used as dynatrons.

by several means. The simplest way is to adjust the plate voltage initially to the point $a$ of zero plate current as shown in Fig. 51. This point is substantially independent of the grid bias so that the negative resistance may be controlled while the plate current is still maintained zero. When the proper negative resistance has been obtained, the circuit under test is short-circuited, after which a 1- or 2-volt increment is added to the plate voltage and the resulting change in plate current read by a microammeter. The negative resistance is then equal to the ratio: (added voltage)/(resulting current increment). Cir-

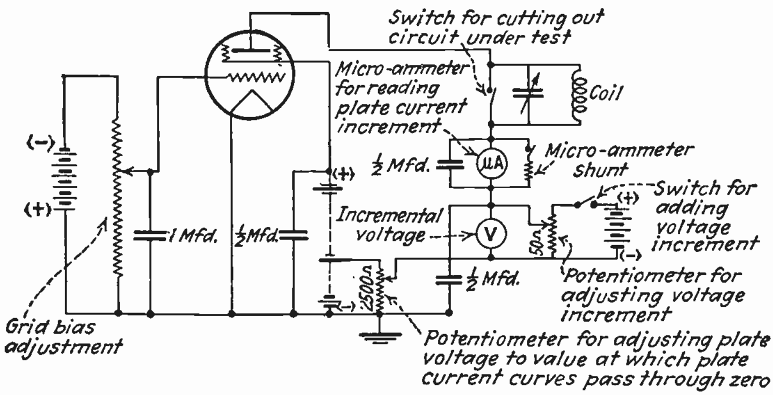



Fig. 52.—Circuit connections for measuring the parallel-resonant impedance of a tuned circuit by the resistance-neutralization method, including means for measuring the negative resistance by the incremental method.

cuit arrangements for carrying out the necessary operations are shown in Fig. 52, where the voltage and current increments are read directly on meters $V$ and $\mu a$, respectively. This method has the disadvantage of being affected by any change in electrode voltages of the tube and by any drift in characteristics, since these cause changes in the d-c plate current that are indistinguishable from those produced by the voltage increment.

*Bridge Methods of Measuring Negative Resistance.*—The most accurate method of measuring the resistance of the dynatron, and the one that is preferable for general use, is to employ a bridge capable of measuring the negative resistance. Suitable circuit arrangements are shown in Fig. 53*a*, together with the equations that give the negative resistance in terms of the bridge arms for the condition of balance.[1] The three circuits shown are equally satisfactory, and the choice between them is determined primarily by constructional considerations. These simple bridge circuits assume that the dynatron offers an impedance which is a pure negative resistance. Actually, the negative plate-cathode resistance is shunted by the plate-cathode interelectrode capacity of the tube, and by stray wiring capacity. As a result it is necessary to modify the circuits, as shown in Fig. 53*b*, in which the capacity in shunt with the negative resistance is balanced by the condenser $C$.[2] The addition of the capacity balance has no effect upon the resistance balance, and adjustments of the two are quite independent.

A complete wiring diagram of equipment that has been found suitable for measuring the resistance of tuned circuits by the resistance-neutralization method is shown in Fig. 54. This includes the dynatron tube, a bridge for measuring the negative resistance of the dynatron, and suitable controls and switches. The arrangement as shown is intended to be operated from a rectifier-filter system. When batteries are to be used, the voltage divider for supplying the plate potential can be omitted and the correct voltage obtained by a tap directly on the battery.

[1] The derivations of these balance equations follow: In the first circuit the current through the telephone receivers will be zero when $R_s$ in parallel with $R_n$ gives an infinite resistance. Since the parallel impedance is $-R_nR_s/(R_s - R_n)$, it is obvious that $R_s = R_n$. In the remaining two circuits the voltage applied between the lower ends of $R_s$ and $R_n$ is

$$\frac{\dfrac{(R_s - R_n)(R - R_1)}{R_s - R_n + R - R_1}}{R_1 + \dfrac{(R_s - R_n)(R - R_1)}{R_s - R_n + R - R_1}}$$

times the oscillator voltage. This multiplied by $(-R_n)/(R_s - R_n)$ gives the voltage between the upper and lower ends of $R_n$, which at balance is equal to the oscillator voltage. This equality is then written in the form of an equation, which when solved for $R_n$ results in the balance equations.

[2] It will be noted that $C_1$, $C$, $R$, and $R_3$ taken by themselves without regard to the remainder of the circuit elements represent a capacity bridge that is balanced by $C$. It will be noted that this bridge is merely superimposed upon the negative resistance bridge of Fig. 53*a* and the two are independent provided $R$ is small.

These circuits were first described by F. E. Terman, Improved Circuits for Measuring Negative Resistance, *Electronics*, vol. 6, p. 340, December, 1933.

The resistance of a tuned circuit is measured by manipulating this set-up in the following manner: The tuned circuit under test is connected in series with the plate circuit at the points $XX$, switch $S_1$ is opened and $S_2$ closed. The latter switch by-passes the negative resistance bridge to alternating currents and so effectively removes it from the circuit. The grid bias is now reduced until the tuned circuit goes into oscillation, and the oscillations are adjusted to the frequency at which it is desired to make measurements. At radio frequencies the existence and also

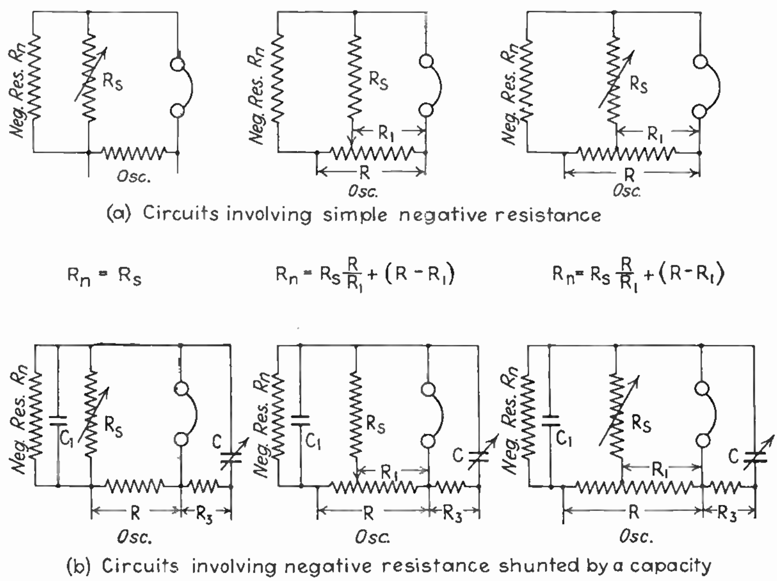



FIG. 53.—Bridge circuits for measuring negative resistance.

the frequency of oscillations can be most easily determined by loose coupling to an oscillating detector that has been calibrated for frequency. Oscillations at audio frequencies can be readily detected by stray leakage to the bridge telephone receivers. The grid bias is then adjusted until the circuit is just on the verge of going in or out of oscillation. Switch $S_1$ is now closed and $S_2$ opened, thus eliminating the circuit being tested and connecting the negative resistance bridge into the plate circuit. It will be noted that in opening $S_2$ the d-c plate current and hence direct-current potentials are not affected. The tube thus continues to operate under the identical conditions that existed with $S_2$

closed. The negative resistance of the dynatron is now balanced by manipulating $R_s$, switch $S_3$, and the balancing condenser. The resistance of the tuned circuit is then $R_s$ if switch $S_3$ is at *a*, and $10R_s$ if it is at *b*. Actually in the latter case the exact value according to the balance equations is $10R_s + 90$, but the 90 ohms is negligible in comparison with $10R_s$ when position *b* must be used.

*Constructional Features of Negative Resistance Bridge.*—Equipment such as illustrated in Fig. 54 is not commercially available,

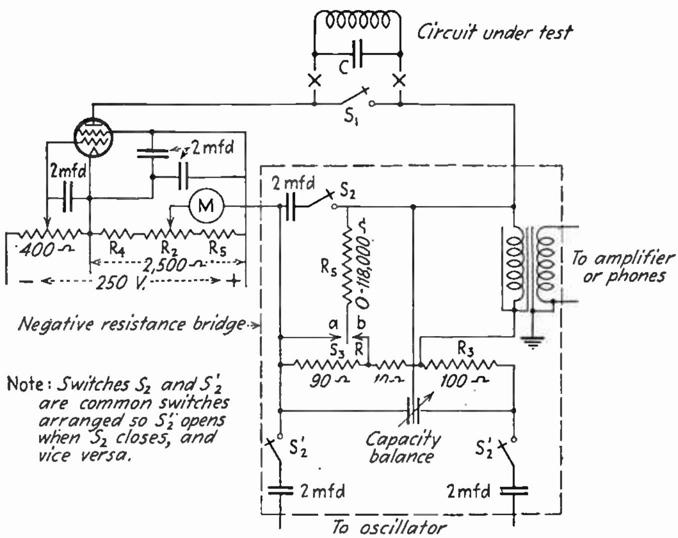



FIG. 54.—Complete circuit diagram of equipment for measuring the parallel-resonant impedance of a tuned circuit by the resistance-neutralization method, using a bridge for determining the negative resistance of the dynatron.

but can be assembled from parts readily obtainable. The grid-bias potentiometer can be of the wire-wound type and should be tapered (unless the tube employed is of the variable-mu type) in order to prevent the adjustment from becoming too critical as cut-off is approached. The resistance $R_s$ can be made up of a single decade resistance giving 100,000 ohms in 10,000-ohm steps, supplemented by a high-grade wire-wound potentiometer of 15,000 to 18,000 ohms with hand-calibrated dial. The resistances $R$ and $R_3$ can be a single layer of resistance wire wound upon a thin strip and tapped at suitable points. The double stator condenser can be made by rebuilding a two-gang midget condenser, or by adding a second stator to an ordinary midget condenser. It is often necessary to connect fixed padding condensers on one side of the condenser to compensate for stray wiring capacities.

The output transformer must be carefully shielded so that spurious currents will not be induced in its secondary. The ideal arrangement is a double electrostatic shield around the primary, as shown in Fig. 54. For comparative measurements, and where precision is not essential, an ordinary audio transformer can be used if care is taken to ground the inner side of the secondary. This causes the first layer of the secondary winding to act as an electrostatic shield, and permits accurate measurements to be made up to at least 100,000 ohms negative resistance, while comparative results may be obtained when the negative resistance is higher. The primary inductance of the output transformer should be designed to operate out of an impedance of about 15,000 ohms, while the turn ratio depends upon the load impedance. Audio-frequency oscillations sometimes occur as a result of the output transformer acting as a tuned circuit, but they can be stopped by shunting a resistance across the transformer.

In setting up the circuit of Fig. 54, care must be taken to see that no spurious couplings exist between the output transformer and the remainder of the circuit, and also to see that the shielding of the output transformer is effective. Spurious couplings to the output system can be tested for by turning off the heater of the tube, disconnecting both terminals of the primary winding from the remainder of the circuit, and grounding the lower side of the primary and its electrostatic shield. When this is done, no sound should be heard in the output system with the oscillator operating at full volume. Spurious couplings may be due to direct coupling between the oscillator or its leads and the output system, or between the oscillator and the power-supply system of the dynatron. Another possible source of trouble is inductive coupling between the transformer and resistances $R$ and $R_3$. The transformer shielding can be tested by disconnecting $R_3$ and leaving the tube filament unlighted. It should then be possible to eliminate the sound in the telephone receivers by adjusting the capacity balance. If this cannot be done, either stray couplings exist or the shielding is not functioning properly.

The accuracy obtainable in measuring parallel-resonant impedance of a tuned circuit should normally be 1 per cent or better. The accuracy can be tested by setting up a circuit reasonant at about 1000 cycles and having an air-core inductance. The impedance of such a circuit can be accurately measured by a general-purpose audio-frequency bridge at the exact resonant frequency, and then remeasured with the resistance-neutralization measuring equipment.

The circuits of Figs. 52 and 54 are suitable for measuring the impedance in resonant circuits from the lowest audio frequencies up to frequencies of 10 to 30 mc. At higher frequencies an uncertainty is introduced by the fact that the time which it takes an electron to travel from cathode to plate is no longer entirely negligible compared with the time represented by a single cycle. The effect of this time of flight is to cause the negative resistance to be different from its value at audio frequencies.[1] The extent of this effect in dynatrons is not known, but results at ultra-high frequencies must be accepted with some reservations.

In making measurements at frequencies such as 5 to 10 mc, it is usually found that the base of the dynatron tube and the tube socket introduce losses which are not negligible compared with the tuned-circuit losses. Errors from this source can be avoided by using an unbased dynatron tube.[2]

The tuning condenser $C$ shown in Fig. 54 can be made an integral part of the dynatron bridge equipment. If calibrated, this condenser can be used for substitution measurements of capacity. Also, by providing suitable coils the equipment can be used as a dynatron wavemeter.

The equipment shown in Fig. 54 will measure impedances from slightly over 1 megohm down to about 15,000 ohms. The lower limit is determined by the dynatron characteristics of the tube, and with commercial tubes will vary from about 8000 to 20,000 ohms. If low impedances are to be measured, a tube having especially low negative resistance must be selected, or two tubes having similar negative resistance characteristics can be paralleled.

[1] See F. B. Llewellyn, Vacuum-tube Electronics at Ultra-high Frequencies, *Proc. I.R.E.*, vol. 21, p. 1532, November, 1933.

[2] Also see H. Iinuma, Resonant Impedance and Effective Series Resistance of High Frequency Parallel Resonant Circuits, *Proc. I.R.E.*, vol. 19, p. 467, March, 1931.

*Resistance-variation Method of Measuring Tuned-circuit Resistance.*—The resistance-variation method of determining the resistance of tuned circuits makes use of the fact that at resonance the current in a circuit is equal to the applied voltage divided by the circuit resistance. If the applied voltage is kept constant, it is then possible to deduce the actual circuit resistance by the current change that results when a known resistance is added to the circuit.

Circuit arrangements suitable for carrying out the necessary measuring operations are shown in Fig. 55. The circuit under test is loosely coupled to a driving oscillator, and has in series with it a thermocouple milliammeter $MA$ and an adjustable

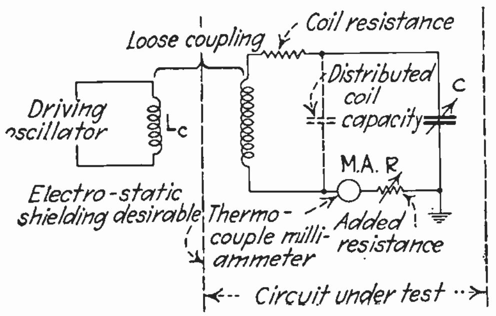



FIG. 55.—Circuit arrangement for measuring radio-frequency resistance by the resistance-variation method.

resistance $R$. The circuit is first tuned to resonance with the driver and the current in the milliammeter is observed when the added resistance $R$ is zero. A known amount of resistance is then added by $R$, the circuit is returned to resonance (if this is necesssary) without changing the coupling to the driver, and the resulting current noted. The apparent series resistance of the circuit is then given by the following formula:[1]

$$\left.\begin{array}{l}\text{Apparent series resistance}\\ \text{of tuned circuit}\end{array}\right\} = R\left(\frac{I_1}{I_0 - I_1}\right) \qquad (22)$$

where $I_0$ and $I_1$ are the currents when the added resistance is respectively zero and $R$. The resistance as measured includes

[1] The derivation of Eq. (22) follows: If $E_0$ is the voltage induced in the circuit and $R_0$ the circuit resistance, then with no added resistance $I_0 = E_0/R_0$ while when the resistance $R$ has been added $I_1 = E_0/(R_0 + R)$. Solving these two equations to eliminate $E_0$ leads to Eq. (22).

the heater resistance of the thermocouple meter, which must be subtracted.

In order to obtain accurate results with this method, it is necessary that the current through the coupling coil $L_c$ be constant, and that the only coupling between the oscillator and the circuit under test be inductive. These requirements can be most satisfactorily met by loose coupling between the two circuits, or by the use of an electrostatic shield. As a check it is always desirable to repeat the measurements with several different values of added resistance. In order to avoid errors from capacities to ground, it is necessary to ground one side of the condenser and place the milliammeter and added resistance on the grounded side of the circuit, as shown in Fig. 55. The added resistance must have negligible skin effect and a reasonably good phase-angle characteristic. It can be either a high-grade decade resistance box or a short link of resistance wire.

Results obtained by the resistance-variation method must be corrected for distributed coil capacity if reasonable accuracy is desired. This is apparent from Fig. 55 where it is seen that the resistance actually measured is the apparent coil resistance, *i.e.*, the resistance effectively in series with the added resistance. In order to obtain the resistance actually in series with the *total* capacity, it is necessary to determine the distributed capacity of the coil when it is in the same situation with respect to the surroundings as in the measuring circuit of Fig. 55. The magnitude of this correction factor is given by rearranging Eq. (8*b*) as follows:

$$\left.\begin{matrix}\text{True series}\\ \text{resistance}\end{matrix}\right\} = \left\{\begin{matrix}\text{apparent series}\\ \text{resistance}\end{matrix}\right\}\left(\frac{C}{C_0 + C}\right)^2 \qquad (23)$$

where $C$ and $C_0$ are the external tuning and distributed capacities, respectively. This correction factor usually differs appreciably from unity, being 0.83 when $C_0/C = 0.1$, and reaching 0.96 when $C_0/C = 0.02$.

*Capacity-variation Method of Measuring the Resistance of Tuned Circuits.*—In this method, the circuit under test is loosely coupled to a driving oscillator of the desired frequency, and the induced current $I_0$ observed at resonance. The tuning capacity of the circuit under test is then increased to some value $C_2$ at which the current has dropped to some convenient value $I_1$, after which the capacity is reduced to a value $C_1$ such that the current is again reduced to the value $I_1$. Then[1]

$$\text{Apparent series resistance} = \frac{C_2 - C_1}{2\omega C_1 C_2}\sqrt{\frac{I_1^2}{I_0^2 - I_1^2}} \qquad (24)$$

where $\omega$ is $2\pi$ times the driver frequency. The resistance as measured includes the heater resistance of the thermocouple meter used to read current.

As with the resistance-variation method, the results obtained by the reactance-variation method of measuring must be corrected for distributed coil capacity by the use of Eq. (23).

The circuit for the reactance-variation method is essentially the same as that used for resistance variation, except that no added resistance need be provided, and the tuning condenser must either be accurately calibrated or, what is still better, shunted by a small calibrated vernier condenser.

The measuring procedure described above is sometimes modified by observing the variation in response on a vacuum-tube voltmeter shunted across the tuning condenser, and omitting the thermocouple. By noting that the voltage across a condenser is $I/\omega C$, Eq. (24) may be modified as follows:

$$\left.\begin{array}{l}\text{Apparent series}\\ \text{resistance}\end{array}\right\} = \frac{C_2 - C_1}{2\omega C_1 C_2}\sqrt{\frac{E_1^2}{(2C_2/C_1 + C_2)^2 E_0^2 - E_1^2}} \qquad (24a)$$

where the notation is as before except that $E_0$ is now the voltage read by the voltmeter at resonance, and $E_1$ is the voltage read by the voltmeter when the tuning capacity is either $C_1$ or $C_2$. For most purposes the formula can be simplified by considering

[1] This equation can be derived as follows: When tuned to resonance,

$$E = I_0 R$$

When detuned, the circuit has a reactance $\Delta X$, so that

$$E = I_1\sqrt{R^2 + \Delta X^2}$$

Solving these two equations simultaneously yields

$$R = \Delta X\sqrt{\frac{I_1^2}{I_0^2 - I_1^2}}$$

Now the difference between the reactances when tuned with $C_1$ and $C_2$ is $2\Delta X$, or $2\Delta X = 1/\omega C_1 - 1/\omega C_2$, and Eq. (24) follows at once.

$2C_2/(C_1 + C_2)$, to be equal to unity. When very precise results are required, or when the measurements are made at very high frequencies, the vacuum-tube voltmeter losses must be determined and allowed for.

*Frequency-variation Method of Measuring the Resistance of Tuned Circuits.*—In this method of measurement the circuit under test is loosely coupled to an oscillator and the response at resonance noted. The frequency of the driving oscillator is then increased to some value $f_1 = \omega_1/2\pi$ at which the response has dropped to a convenient reference point, after which the frequency is increased beyond resonance to a value $f_2 = \omega_2/2\pi$ at which the response is the same as at $f_1$. *During this process the current in the coil of the driving oscillator must be kept constant.*

If the response is measured by a vacuum-tube voltmeter across the tuning condenser, then it can be shown that with inductive coupling

$$\left.\begin{array}{l}\text{Actual series resistance}\\ \text{of circuit}\end{array}\right\} = 2\pi L(f_2 - f_1)\sqrt{\frac{E_1^2}{E_0^2 - E_1^2}} \qquad (25a)$$

$$Q \text{ of circuit} = \frac{\dfrac{f_0}{f_2 - f_1}}{\sqrt{\dfrac{E_1^2}{E_0^2 - E_1^2}}} \qquad (25b)$$

$$= \frac{\dfrac{\sqrt{f_1 f_2}}{f_2 - f_1}}{\sqrt{\dfrac{E_1^2}{E_0^2 - E_1^2}}} \qquad (25c)$$

where $f_2$ and $f_1$ are the frequencies above and below resonance for which the voltage read by the vacuum-tube voltmeter is $E_1$, $f_0$ is the frequency at resonance, where the voltage is $E_0$, and $L$ is the actual inductance of the circuit.

If the response is measured by a thermocouple meter in the tuned circuit, Eq. (25*b*) can be modified by noting that $E = I/\omega C$, which leads to

$$\left.\begin{array}{l}\text{Actual series resistance}\\ \text{of circuit}\end{array}\right\} = 2\pi L(f_2 - f_1)\sqrt{\frac{I_1^2}{(f_2/f_1)I_0^2 - I_1^2}} \qquad (25d)$$

where $I_0$ and $I_1$ are respectively the currents at resonance, and at $f_1$ and $f_2$, and the remainder of the notation is as above.

It will be observed that the frequency-variation method gives the actual rather than the apparent circuit resistance. When a vacuum-tube voltmeter is used, allowance must be made for the input losses of the tube if there are any. If a thermocouple meter is used, the subtraction necessary to correct for meter resistance involves the distributed capacity because the meter is in series with only part of the total capacity. If the heater resistance of the instrument is $R_m$, the equivalent series resistance to be subtracted from Eq. (25*d*) is then $R_m\ (C/C + C_0)^2$ where $C$ and $C_0$ are respectively external tuning and distributed capacities.

The accuracy of the frequency-variation method is determined largely by the precision with which $(f_2 - f_1)$ is known. This can be made very great by measuring the difference directly as discussed in Sec. 35 instead of determining $f_2$ and $f_1$ individually and then subtracting to obtain their small difference.

Modifications in the measuring technique outlined above may be desirable at ultra-high frequencies because of vacuum-tube voltmeter losses resulting from the fact that the time of flight of electrons within the tube is not negligible, or because skin effect makes the losses in a thermocouple heater uncertain. These difficulties can be minimized by coupling the indicating device very loosely to the circuit under test. Thus the vacuum-tube voltmeter may be connected across only a single turn of the inductance coil, or placed across a large condenser in series with the tuning condenser. Such arrangements usually require increased power in the driving oscillator.

*Relative Merit of Methods for Measuring the Resistance of Tuned Circuits.*—The choice of a method of measurement from among the several possibilities depends largely upon the equipment available, the precision required, etc. The resistance-neutralization and frequency-variation methods are best, since they require no correction for distributed capacity of the coil. The resistance-neutralization method is to be preferred for measurements at all except the highest frequencies when a completed set-up such as shown in Fig. 54 is available. The frequency-variation method is the only accurate method available at ultra-high frequencies and is also best for other frequencies when apparatus as illustrated in Fig. 54 is not available.

The resistance- and capacity-variation methods have the very important drawback that their results must be corrected for distributed capacity. The capacity-variation method has the further disadvantage that it is difficult to measure with accuracy the small changes in capacity involved.

*Separation of Losses.*—The foregoing methods of measuring circuit resistance give the total resistance of the circuit and say nothing as to how this total is distributed between the coil and condenser. When well-constructed air-dielectric condensers are used for tuning, the condenser losses are extremely low, so that for most practical purposes it is permissible to assume that the coil is responsible for the entire circuit loss.

The condenser losses can be evaluated, however, with the aid of a calibrated variable condenser designed according to the method described in Sec. 25 so that the losses are independent of the capacity setting.[1] With such a condenser, the procedure for measuring the series resistance of a condenser is as follows: The condenser under investigation and the specially designed variable condenser are connected in parallel and the combination used as the tuning condenser of a resonant circuit. The series resistance of the circuit is measured by any means available. The condenser under test is then removed, and the standard condenser readjusted until the same resonant frequency is obtained as previously, after which the series resistance of the circuit is again measured. The difference between the two results is then related to the equivalent series resistance of the condenser under test by one of the following equations:[2]

[1] C. T. Burke, Substitution Method for the Determination of Resistance of Inductors and Capacitors at Radio Frequencies, *Trans. A.I.E.E.*, vol. 46, p. 482, 1927.

[2] Equation (26*a*) is derived as follows: The current through $C_x$ is $(C_x/C_0)$ times the current through $C_0$ when the voltage across the condensers is kept constant. Hence a resistance $R$ in series with condenser $C_x$ has the same losses as a resistance $\Delta R$ in series with the capacity $C_0$ when $\Delta R$ and $R$ satisfy the relation

$$\frac{R}{\Delta R} = \left(\frac{C_0}{C_x}\right)^2$$

In Eq. (26*b*) the difference between the resistances $R_1$ and $R_2$ is accounted for by the equivalent shunt resistance $R_{sh}$ of $C_x$, which is placed in parallel with $R_2$ to give $R_1$. Hence

$$\frac{R_2 R_{sh}}{R_2 + R_{sh}} = R_1$$

The first form of Eq. (26*b*) follows by solving for $R_{sh}$, and the second form results when one replaces the shunt resistance by a series resistance giving the same power factor.

$$\left.\begin{array}{l}\text{Series resistance of}\\ \text{condenser}\end{array}\right\} = \Delta R\left(\frac{C_0}{C_x}\right)^2 \tag{26a}$$

where $\Delta R$ is the change in *apparent* series resistance of the circuit, $C_x$ is the capacity of the condenser being tested (which is equal to the change in capacity of the calibrated condenser required to restore the resonant frequency), and $C_0$ is the capacity which the calibrated condenser supplies to tune the circuit after the removal of $C_x$. An alternative form of Eq. (26*a*), suitable for use when the equivalent parallel-resonant impedance of the tuned circuit is determined, is

$$\left.\begin{array}{l}\left.\begin{array}{l}\text{Equivalent shunt resistance}\\ \text{of condenser}\end{array}\right\} = \dfrac{R_2R_1}{R_2 - R_1}\\ \left.\begin{array}{l}\text{Equivalent series resistance}\\ \text{of condenser}\end{array}\right\} = \dfrac{R_2 - R_1}{(\omega C_x)^2R_2R_1}\end{array}\right\} \tag{26b}$$

where $R_1$ and $R_2$ are the parallel resonant resistances with and without $C_x$ in the circuit, respectively, and $\omega$ is $2\pi$ times the resonant frequency of the circuit. Equation (26*a*) is the proper equation to use when circuit losses are determined by the resistance-variation or reactance-variation method, while Eq. (26*b*) is especially adapted for use with the resistance-neutralization method of measuring circuit losses.

Several other means have been devised for determining the equivalent series resistance of a condenser. Thus the power loss can be measured directly by use of a calorimeter, but the technique is slow and cumbersome. Moullin has developed a method which employs several coils that are identical except for wire resistivity, and are so proportioned that the *relative* resistances can be calculated. These coils are connected in turn to the condenser being tested, and the series resistance of the circuit measured for each case. The results make possible the calculation of condenser series resistance.[1]

**22. Radio-frequency Bridges.**—The simplicity and convenience of bridge measurements have led to the development of bridges capable of operating at radio frequencies. Such bridges must be very carefully built because at high frequencies even small stray capacities have low reactances, while small inductances have large

[1] An example of this method is described by W. Jackson, The Analysis of Air Condenser Loss Resistance, *Proc. I.R.E.*, vol. 22, p. 957, August, 1934.

reactances. The result is that circuit elements which can be ignored at 1000 cycles may become the controlling factor at 1,000,000 cycles and introduce large errors.

Several forms of successful radio-frequency bridges have been devised.[1] These are all simple unity-radio bridges shielded in such a way as to control stray capacities and localize their effects in a definite manner. A completely shielded radio-frequency bridge is shown in Fig. 56 and gives one solution of the problem involved.

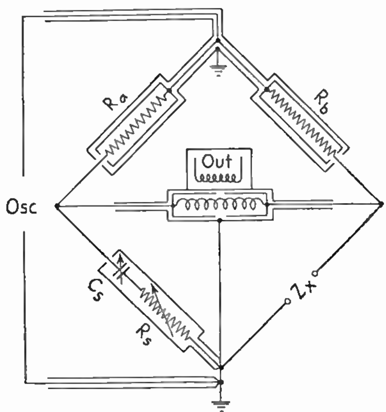



Fig. 56.—Typical completely shielded radio-frequency bridge.

The ratio arms $R_a$ and $R_b$ are made of non-inductive, low-capacity resistance units mounted in individual shields, with these in turn surrounded by a grounded shield. By making everything perfectly symmetrical it is possible to equalize the shunting capacities and make the ratio exactly unity. The arrangement shown for the driving circuit, with one terminal grounded, makes an input transformer unnecessary, thus avoiding the possibility of coupling to the output transformer. The primary of the output transformer is provided with a double shielding system. The two halves of the primary winding, as well as the shields, must be exactly identical as far as capacitive couplings are concerned, since the output transformer introduces capacities that are in shunt with the unknown and standard arms of the bridge. The bridge is shown provided with condenser $C_s$ and resistance $R_s$ is the standard arm.

[1] For a further discussion of radio-frequency bridges including detailed descriptions of several forms of such bridges, see John G. Ferguson, Shielding in High Frequency Measurements, *Trans. A.I.E.E.*, vol. 48, p. 1286, October, 1929; W. J. Shackelton, A Shielded Bridge for Inductive Impedance Measurements, *Trans. A.I.E.E.*, vol. 45, p. 1266, 1926; Charles T. Burke, Bridge Methods of Measuring at Radio Frequencies, *General Radio Experimenter*, vol. 7, p. 1, July, 1932; Leo Behr and A. J. Williams, Jr., The Campbell-Shackelton Shielded Ratio Box, *Proc. I.R.E.*, vol. 20, p. 969, June, 1932.

Balance in a radio-frequency bridge can be detected by an oscillating radio receiver if the driving voltage is unmodulated, or by an ordinary receiver if the driving voltage is modulated. In using the bridge, care must be taken to avoid direct pick-up of the oscillator by the receiver.

The accuracy obtainable depends upon conditions. The best results are obtained with the substitution method, which can yield accuracies better than 1 per cent at frequencies exceeding one million cycles. Errors several times as great as this can be expected with direct measurements. In every case care must be taken in arranging the leads associated with the unknown and standard impedances, since the leads introduce impedances and admittances that are by no means negligible.

# CHAPTER IV

## RESISTANCE, INDUCTANCE, AND CAPACITY DEVICES

**23. Resistance Devices.** *Resistance Wire.*—The resistance alloys most commonly used in resistors for communication purposes are nichrome and related nickel-chromium alloys such as chromel, advance and similar copper-nickel alloys, and manganin. The principal properties of these materials are given in the accompanying table.

TABLE III.—PROPERTIES OF RESISTANCE WIRE

| Properties of material | Resistivity: Ohms per mil foot | Resistivity: Temperature coefficient per °C. | Thermal e.m.f. against copper, $\mu$v/°C. | Maximum working temperature, °C. |
|---|---|---|---|---|
| Nichrome, nichrome I–V, chromel C, etc. | 650 | 0.0002 | 22 | 1000 |
| Advance, constantan, $I_a$–$I_a$, copel, ideal, etc. | 295 | Nil | 43 | 450 |
| Manganin | 290 | Less than 0.00001 | 2 | |

NOTE.—The foregoing are average values. Wire made by different manufacturers and even different lots from the same manufacturer may differ slightly in characteristics.

The nichrome group is capable of operating at high temperatures and is the wire generally employed in rheostats, resistance tubes, etc. It will also be noted that nichrome has a resistivity over twice as great as the other resistance materials and its temperature coefficient, although not zero, is small, being only one-twentieth that of copper. Manganin is always used in the construction of precision resistances because of its stability and negligible temperature coefficient and freedom from thermoelectric effects with copper. Advance can often be used interchangeably with manganin as both have negligible temperature coefficient and similar resistivities. Advance possesses a high thermoelectric coefficient against copper, however, and so is not suitable for some applications, such as shunts for direct-current instruments.

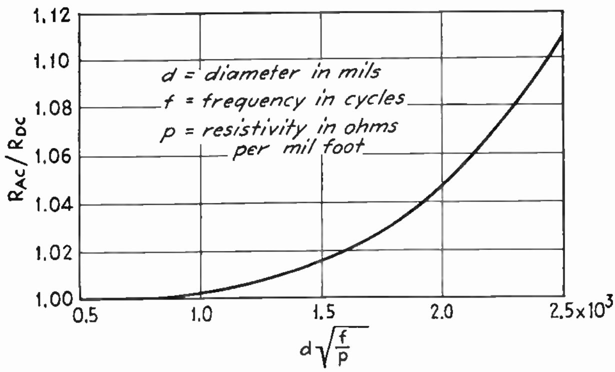



FIG. 57.—Skin-effect resistance ratio as a function of frequency and wire characteristics.

When stability of characteristics is important, it is necessary to anneal the wire after it has been wound in place in order to remove mechanical strains which would otherwise produce a change of resistance with age and make the temperature coefficient highly erratic. This annealing can be accomplished by baking at a temperature of 120°C. for 24 hr.

TABLE IV.—LARGEST PERMISSIBLE WIRE DIAMETER IN MILS FOR SKIN-EFFECT RATIO OF 1.01

| Frequency, kilocycles | Nichrome | Advance and manganin |
|---|---|---|
| 100 | 104.5 | 70.2 |
| 200 | 74.0 | 49.6 |
| 500 | 46.8 | 31.4 |
| 1000 | 33.1 | 22.2 |
| 1500 | 27.0 | 18.1 |
| 2000 | 23.4 | 15.7 |
| 3000 | 19.1 | 12.8 |

In resistances used at high frequencies, attention must be given to skin effect. The ratio of alternating-current to direct-current resistance of an isolated wire is determined by the factor $d\sqrt{f/\rho}$, where $d$ is the wire diameter, $f$ the frequency, and $\rho$ the resistivity. The relation for low resistance ratios is given in Fig. 57,[1] which can be used to determine the largest wire permissible at any given frequency. Thus the accompanying table derived from Fig. 57 gives the largest wire that can be used at various frequencies and still keep the difference between the alternating-current and direct-current resistance within 1 per cent.

*Phase Angles of Resistance Units.*—Every resistance unit can be represented by the equivalent circuit shown in Fig. 58, in which $R$ is the actual resistance introduced by the wire, $L$ is the inductance produced by the magnetic flux associated with the current in the wire, and $C$ is the distributed capacity plus capacity between terminals and connecting wires. The effect of this inductance and capacity is to make the impedance of the resistance, as measured at its terminals, have a slight leading or lagging component which depends upon the construction of the resistance, and in any particular case varies with frequency. The amount by which the power-factor angle of the resistance departs from unity is termed the phase angle and is a measure of the reactive effects associated with the resistance.

Analysis of the circuit of Fig. 58 shows that when the phase angle is small, as is always the case under conditions for which the resistance is used, the resistance acts as though it had in series with it an equivalent inductance $L_{eq} = L - R^2C$. It will be noted that this equivalent inductance is independent of frequency and may be either positive or negative (capacitive) depending upon the resistance. It is usually inductive for low resistances and capacitive for high resistances with the transition when the resistance is of the order of hundreds of ohms. The phase angle is related to the equivalent inductance by the equation

$$\text{Phase angle in radians} = \frac{\omega L_{eq}}{R} = \frac{\omega(L - R^2C)}{R} \qquad (27)$$

The phase angle is seen to be proportional to frequency. The merit of a resistance unit from the point of view of low phase angle is obviously inversely proportional to the time constant $L_{eq}/R$.

[1] This has been calculated from the usual formulas for skin effect found in electrical engineers' handbooks.

*Characteristics of Commercial Resistances.*—The wire-wound rheostats, potentiometers, and fixed resistances used in communication equipment nearly always employ nichrome or similar wire in a single-layer winding. While such resistance devices are designed primarily on the basis of current-carrying capacity, they, nevertheless, usually have reasonably low phase angles

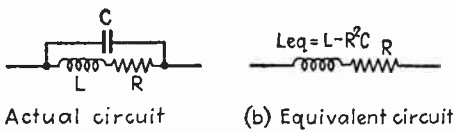



FIG. 58.—Actual and equivalent circuit of a resistance, showing associated inductance and capacity.

at audio frequencies. Some typical characteristics are given in the following table:

TABLE V

| Description | Resistance, ohms | Phase angle at 1000 cycles |
|---|---|---|
| 1 watt wire wound. . . . . . . . . . . | 3,500 | 0°1′ lag |
| 1 watt wire wound. . . . . . . . . . . | 18,000 | 0°6′ lead |
| 20 watt wire wound. . . . . . . . . . | 50,000 | 0° |
| 200-watt voltage divider. . . . . . | 100,000 | 2°54′ lead |
| 50-watt voltage divider. . . . . . . | 26,000 | 0°12′ lag |
| Service test box. . . . . . . . . . . . . | 50,000 | 1°40′ lead |
| Small rheostat. . . . . . . . . . . . . . . | 5,000 | 0°4′ lag |
| Potentiometer. . . . . . . . . . . . . . . | 50,000 | 0°24′ lead |

It will be noted that the phase angle, although varying widely according to the method of construction, is in all cases so small at audio frequencies that the reactances associated with the resistance have negligible effect upon the impedance. These commercial resistance devices are therefore satisfactory for many audio-frequency applications such as voltmeter multipliers, attenuators, and voltage dividers. It will be noted that when the resistance is low the impedance tends to be inductive, while high-resistance units have a capacitive phase angle.

The carbon resistors commonly used in radio receivers, and also the metal-film type of resistor, possess negligible inductance and capacity effects. The carbon resistors are not stable, however, and their resistance changes considerably with temperature, applied voltage, and age. Pencil leads, which are sometimes found useful as non-inductive resistances in experimental work are fundamentally similar to the commercial carbon resistors. The metal-film types can be made much more stable and in fact are the only available standards for very high resistances, such as 100 megohms.

*Resistors Having Low Capacitive and Inductive Effects.*—The inductance of a resistance is determined primarily by the number

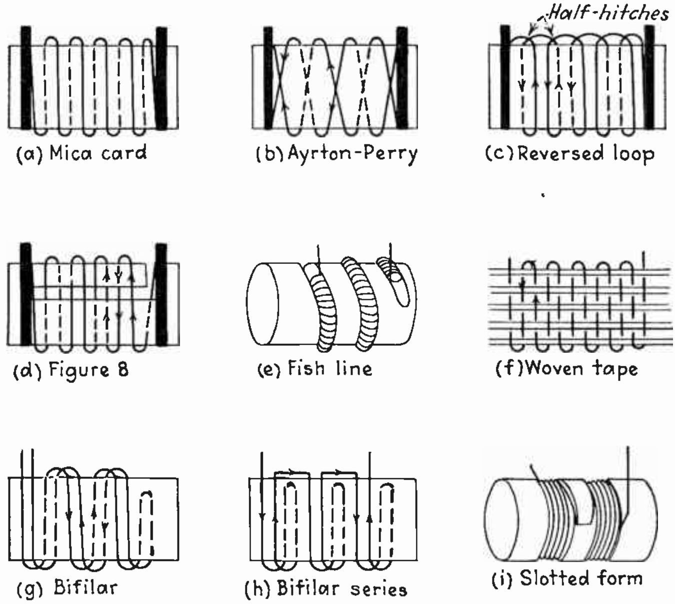



FIG. 59.—Types of resistance windings used to minimize reactive effects.

of turns of wire and the area inclosed by the individual turns. To keep the inductance low, each turn should inclose the minimum possible area, and the wire should have as many ohms per foot of length as possible so that the length required to obtain the desired resistance will be small. In addition, it is desirable that adjacent turns carry current in opposite directions so that the residual inductance of an individual turn is neutralized by the effect of adjacent turns. A low capacitive reactance associated with a resistance is obtained by arranging the winding in such a way that adjacent turns of wire have a low potential difference between them and are as far apart as possible.

Methods that can be used to minimize the reactive effects associated with a resistance are shown in Fig. 59. The mica-card type of resistance uses a single-layer winding on a thin mica form provided with copper end strips that serve as terminals and reinforcing. A low inductance can be obtained by making the card very thin and using small wire to give a high resistance per turn.[1] The Ayrton-Perry type of resistance is constructed by winding a single spaced layer of wire on a thin strip; after which a second wire is wound in opposite direction between turns of the first winding. The two windings are connected in parallel and thereby produce practically zero resultant magnetic effect. The distributed capacity is low because adjacent turns have very little potential difference between them. The reversed-loop winding obtains low inductive effects by making a half hitch at the end of each turn, and thus reversing the direction of current in adjacent turns. The winding of Fig. 59*d* accomplishes substantially the same result in a different way. The fish-line type of resistance consists of a fine resistance wire wound over a silk cord that serves as a core, and the resulting "fish line" is then space wound on a cylindrical form. The tape resistance

[1] Resistance units wound on thin cards are the easiest to construct, so that it is important to be able to determine their limitations. If the form is thin compared with the axial length of the winding, the inductance is given by the formula

$$\left.\begin{array}{r}\text{Series inductance}\\ \text{of resistance card}\end{array}\right\} = 0.032AbT^2 \quad \mu\text{h}$$

where $A$ is the area inclosed by an individual turn in square inches (measurements being made to the center of the wire), $b$ the axial length of the winding in inches, and $T$ the turns per linear inch. If the form does not entirely meet the requirement of being very thin compared with the axial length, the inductance calculated by the foregoing formula is greater than the inductance that will actually exist. Substitution of numerical values in the equation for inductance shows that resistances of a few ohms or more which have entirely satisfactory characteristics at audio and the lower radio frequencies can be wound on cards of thin bakelite provided *small* wire is used. Thus a 1000-ohm winding with No. 43 manganin on a bakelite card 2 in. wide and $\frac{1}{32}$ in. thick, if wound 100 turns to the inch, will have an axial length of about 0.5 in. The inductance will be 10 $\mu$h, and in the absence of stray shunting capacities the impedance at 1000 kc will be in error by only about $\frac{1}{5}$ per cent. Before they are wound the bakelite cards should be baked for 24 hr. at 120°C. to stabilize their dimensions and so prevent the wire loosening upon annealing.

is made by weaving the resistance wire into a fabric in which the wire serves as the woof while silk thread functions as the warp.[1] The bifilar winding has negligible inductance but the capacity is relatively large because the beginning and end of the resistance are close together. This capacity effect is minimized to some extent by subdividing the total resistance into several bifilar sections as shown in Fig. 59*h*. The slotted-form winding will have moderately low capacity because of the subdivision of the winding, and the inductance can be kept moderately low by reversing the direction of the winding in adjacent slots and by using small wire to keep down the required size and the number of turns.

The mica-card, reversed-loop, and figure-eight types of resistances can be made to have very low phase angles and are the types used in radio-frequency attenuators. The mica-card, fish-line, and woven-tape types of construction are commonly used in decade resistance boxes designed to have a low phase angle at radio frequencies, particularly for the high-resistance units. The Ayrton-Perry winding is used in one make of decade box for 1-, 10-, and 100-ohm units, while the simple bifilar is used by the same concern for 0.1-ohm units. The slotted type of construction is used in very high-resistance units (such as meter multipliers) where only moderately low phase angle is essential.

*Resistance Boxes.*—A number of individual resistance units may be mounted in a box and provided with suitable switches for varying the resistance. Possible switching arrangements are shown in Fig. 60. In the circuit of Fig. 60*a*, which is widely used, the resistance units are grouped in decades with each decade controlled by an 11-point switch, thus giving overlap between dials. The arrangement of Fig. 60*b* will accomplish the same result for one decade as Fig. 60*a* with fewer resistances, but it requires a two-gang switch and a complicated wiring system which introduces extra shunting capacity. The circuit of Fig. 60*c* is the most economical of resistance units but is not particularly convenient to manipulate. In this arrangement all the resistances are connected in series, and those not desired are

[1] For a discussion of various types of weaves that may be used, see L. Behr and R. E. Tarpley, Design of Resistors for Precise High-frequency Measurements, *Proc. I.R.E.*, vol. 20, p. 1101, July, 1932.

short-circuited out by inserting plugs at the appropriate places.[1]

The phase angle of resistance boxes depends upon the characteristics of the individual resistance units and upon the wiring, the shielding, and the capacities introduced by the switches. When it is desired to keep the phase angle small, the resistance

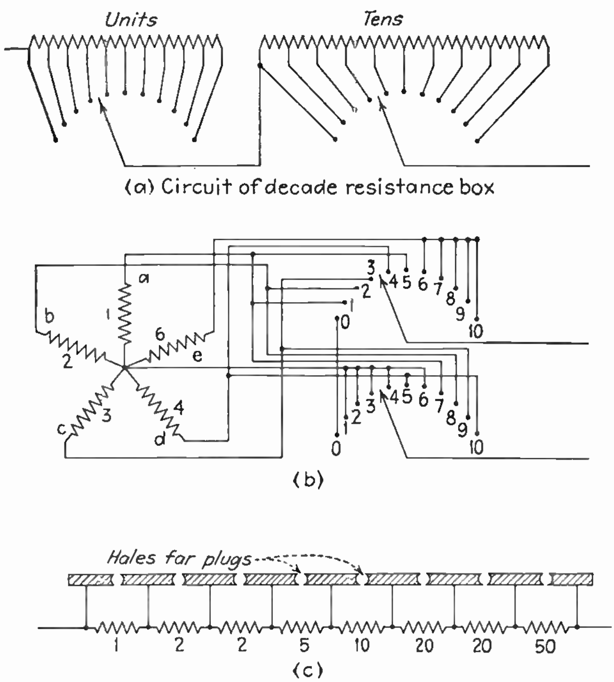



Fig. 60.—Switching arrangements for resistance boxes.

units must be of one of the non-inductive low-capacity types shown in Fig. 59. The wiring must be as simple as possible and the switch should be made as compact as is consistent with low contact resistance. Even with the utmost precautions it is always found that the inductance introduced by the connecting wires, and the extra shunting capacity produced by the leads and switch, will make the phase angle of the resistance box much larger than that of the individual resistance units.

The type of performance that can be expected from a well-made decade resistance box is illustrated by the data in the accompany-

[1] Other switching arrangements in which the units are rotated are described by L. Behr and R. E. Tarpley, *op. cit.*

ing table. Percentages are for maximum setting of each decade. The best resistance boxes have enclosed switches to protect the contacts from dust and are mounted in electrostatically shielded containers. In this way the resistance units and the switch

Characteristics of Commercial Decade Resistance Box[1]

| Decade, ohms per step | Type of winding | Percentage error in resistance | | | Percentage error in impedance | | |
|---|---|---|---|---|---|---|---|
| | | Frequency, kc | | | Frequency, kc | | |
| | | 100 | 1000 | 5000 | 100 | 1000 | 5000 |
| 0.1 | Bifilar | 0.1 | 5 | .... | 0.7 | | |
| 1.0 | Ayrton-Perry | 0 | 1 | 25 | 0.2 | 20 | |
| 10 | Ayrton-Perry | 0 | 0.5 | 11 | 0 | 2 | |
| 100 | Ayrton-Perry | 0 | 0.3 | 4 | 0 | 0.3 | 15 |
| 1,000 | Mica card | 0 | −1 | −30 | 0.1 | 6 | |
| 10,000 | Mica card | −0.2 | −16 | .... | 20 | | |

[1] From *General Radio Experimenter*, February, 1932.

are protected from electrostatic couplings with neighboring objects. The shield may be connected to either terminal of the resistance box or may be left floating. The phase angle will vary depending upon how the shield is handled.[1]

The steps in decade resistance boxes are seldom made less than 1/10 ohm, and sometimes not less than 1 ohm. Finer adjustment of resistance is obtainable by use of a slide-wire arrangement consisting of a single wire mounted on a rotating disk and pressing against a fixed slider. The inductive effect in such a slide-wire arrangement can be minimized by arranging the slider so that it makes contact between the resistance wire and a return path consisting of a copper strip mounted just under the slide wire but insulated from it.

*Attenuators.*[2]—Attenuators are resistance networks used for the purpose of reducing power, voltage, or current in controllable

[1] For a detailed discussion of the problem of shielding resistance units see J. G. Ferguson, Shielding in High Frequency Measurements, *Trans. A.I.E.E.*, vol. 48, p. 1286, October, 1929.

[2] A more complete discussion of the design of attenuators is given by P. K. McElroy, Designing Resistive Attenuating Networks, *Proc. I. R. E.*, vol. 23, p. 213, March, 1935.

and known amounts. A number of typical attenuator networks are shown in Figs. 61 and 63. These may be classified as balanced and unbalanced, and also symmetrical and unsymmetrical types. A balanced attenuator is the same with respect to the two sides of the circuit and is exemplified by the H attenuator of Fig. 61*a*. A symmetrical attenuator is symmetrical with respect to generator and load sides, while a non-symmetrical attenuator is not. The H and T attenuators of Fig. 61 are hence symmetrical, while the remaining attenuators are non-symmetrical.

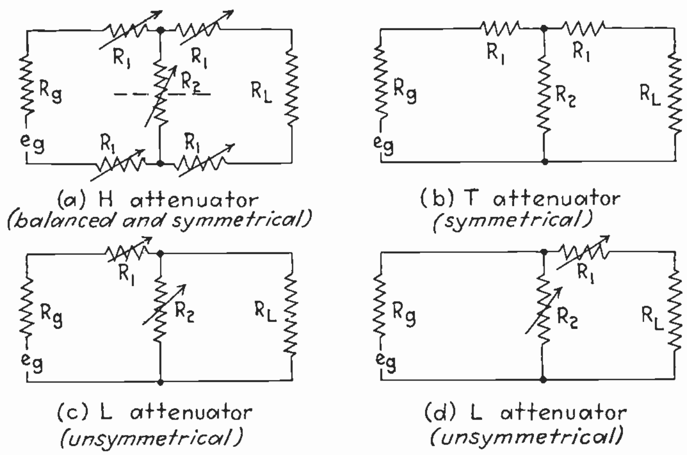



FIG. 61.—Circuit arrangements of T, H, and L attenuators. The resistances $R_L$ and $R_g$ are respectively load and generator resistances.

The important properties of an attenuator are the input and output resistances and the attenuation. The input resistance is defined as the impedance looking into the attenuator from the generator side (but with generator disconnected), when the load resistance $R_L$ is connected. The input resistance represents the load impedance into which the generator operates, and depends upon the load resistance $R_L$ and the attenuator proportions. The output resistance is similarly the impedance which is seen looking into the attenuator from the load side (but with the load disconnected) when the generator resistance $R_g$ is present and the generator voltage $e_g$ is short-circuited. The output resistance represents the equivalent internal impedance of the generator that may be considered as supplying power to the load according to Thévenin's theorem, and depends upon the attenuator proportions and the generator resistance $R_g$. The attenuation can be defined as the insertion loss, *i.e.*, the ratio of the output actually obtained compared with the output that would be obtained if the attenuator were removed. This attenuation can be expressed as the ratio $\alpha$ by which voltage is reduced, or in power loss (decibels). The relation between the two methods of stating attenuation is

$$\text{Loss in decibels} = 20 \log_{10} \alpha \tag{28}$$

The attenuation depends upon the attenuator proportions, and upon the generator and load resistances $R_g$ and $R_L$ in relation to the input and output resistances of the attenuator.

The T attenuator is normally used under conditions where the generator and load resistances are equal, and is then designed so that the attenuator input and output resistances have this same value. Under such conditions the insertion of the attenuator does not alter the impedance relations in the circuit. The design equations of the attenuator, when load, generator, input, and output resistances all have the same value $R_L$, are

$$\begin{aligned} R_1 &= R_L\left(\frac{\alpha - 1}{\alpha + 1}\right) \\ R_2 &= R_L\left(\frac{2\alpha}{\alpha^2 - 1}\right) \end{aligned} \tag{29}$$

where $R_1$ and $R_2$ are the attenuator resistances as shown in Fig. 61*b*, and $\alpha$ is the ratio $E_2/E_1$, where $E_1$ is the voltage actually across the load and $E_2$ is the voltage that would be across the load if the attenuator was removed. It will be noted that, for the matched impedance conditions assumed, $E_2$ is the voltage across the attenuator input terminals.

When a T attenuator such as just described is used with the wrong load resistance (but the proper generator resistance), the input resistance of the attenuator is altered, and an additional attenuation, called a reflection loss, is introduced in addition to the attenuation that is present with the proper load resistance. The reflection loss is given by the relation

$$\left.\begin{array}{l}\text{Added attenuation in decibels} \\ \text{caused by mis-match}\end{array}\right\} = 10 \log_{10} \frac{(1 + r)^2}{4r} \tag{30}$$

where $r$ is the ratio of actual load resistance to the load resistance that matches the attenuator output.

If the generator impedance does not equal the input resistance of the attenuator, when the proper load resistance is used, the effect is to alter the output impedance, and also to make the power delivered to the load less than it would be if an ideal transformer were used to match the generator to the attenuator input resistance. This loss is also called a reflection loss and is given by Eq. (30) if $r$ is taken as the ratio of actual generator resistance to the input resistance of the attenuator. When both generator and load resistances fail to match the attenuator, both input and output impedances are altered, and the additional attenuation cannot be given by a simple expression.

The H attenuator can be thought of as two T attenuators having a neutral point shown by the dotted lines in Fig. 61*a*. The H attenuator is accordingly designed as though it consisted of two similar T attenuators, each having a load resistance equal to half of the actual load resistance and having the same attenuation as is desired from the actual H attenuator.

The L attenuator, as illustrated in Figs. 61*c* and 61*d*, is unsymmetrical and can offer constant impedance from only one side. The impedance as viewed from the other side of the attenuator will depend on the amount of attenuation. When the load and generator resistances are equal, the design formulas for the case where the input resistance of the attenuator of Fig. 61*c*, or the output resistance in Fig. 61*d* is equal to the load resistance, are

$$\left.\begin{aligned} R_1 &= R_L\left(\frac{\alpha - 1}{\alpha}\right) \\ R_2 &= \frac{R_L}{\alpha - 1} \end{aligned}\right\} \tag{31a}$$

$$\left.\begin{array}{l}\text{Resistance on variable imped-}\\ \text{ance side of attenuator}\end{array}\right\} = R_L\left(\frac{(2\alpha - 1)}{\alpha^2 + (\alpha - 1)^2}\right) \tag{31b}$$

If, on the other hand, the load and generator resistances are equal and the output resistance in Fig. 61*c* or the input resistance in Fig. 61*d* is to equal the load resistance, then

$$\left.\begin{aligned} R_1 &= R_L(\alpha - 1) \\ R_2 &= R_L\left(\frac{\alpha}{\alpha - 1}\right) \end{aligned}\right\} \tag{32}$$

$$\left.\begin{array}{l}\text{Resistance on variable impedance}\\ \text{side of attenuator}\end{array}\right\} = R_L\left(\frac{2\alpha^2}{2\alpha - 1}\right) \tag{33}$$

In the L attenuator, failure of the terminating circuit to match the resistance on the side of the attenuator designed for constant resistance alters the impedance at the other side of the attenuator, and also introduces a reflection loss that can be calculated exactly as in the case of the T attenuator, by using Eq. (30). Failure to use the design value of terminating resistance at the variable impedance side of an L attenuator will alter the attenuator resistance at the other end, and will at the same time change the attenuation by an amount

$$\left.\begin{array}{l}\text{Added attenuation in decibels}\\ \text{caused by improper resistance}\end{array}\right\} = 10 \log_{10}\left(\frac{1+\gamma}{1+r}\right)^2 \frac{r}{\gamma} \tag{34}$$

where $r$ is the ratio of design resistance to attenuator resistance on the variable impedance side, as given by Eq. (31*b*) or Eq. (33), and $\gamma$ is the corresponding ratio of the actual terminal resistance which is used instead of the design value, to the attenuator resistance on the variable impedance side. If this added attenuation is negative, as is often the case, it means that the use of the improper resistance reduces the attenuation as compared with the attenuation for the design conditions.

The attenuation introduced by an L, T, or H section can be controlled in discrete steps by varying the resistance branches of the attenuator with tap switches operated from a common shaft. Continuous variation can be obtained by constructing the resistance branches in the form of slide-wire rheostats operated from a common shaft. The winding form of each rheostat must be so tapered that the resistances of the various arms maintain the proper relation to each other as the shaft is rotated to vary the attenuation. The attenuation can also be varied by switching attenuator sections in and out of the circuit as shown in Fig. 62. In such an arrangement each section is designed to have an input resistance equal to the load resistance so that the input resistance of each section provides the correct load resistance for the next preceding section. The individual sections can then be switched in and out of the circuit without affecting the impedance relations, and the total attenuation is the sum of the decibel attenuation of the sections in use (or the product of their voltage ratios). Attenuators greater than 30 to 40 db per section require resistance arms that contain either excessively low or high resistance, and so are most conveniently obtained by the use of several sections of moderate attenuation connected in tandem. One of these sections may be adjustable either continuously or in small steps, to provide fine control of the attenuation, while coarse adjustment is obtained by switching fixed sections in and out of the circuit (see Fig. 62c).

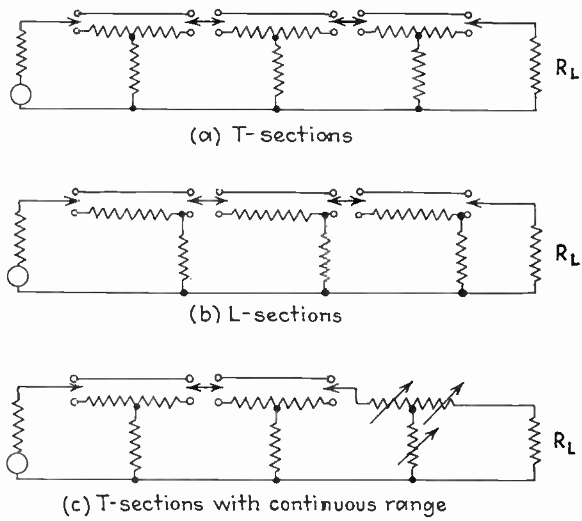



FIG. 62.—Methods of controlling attenuation by switching attenuator sections. The input resistance of each section serves as the load resistance for the preceding section, and the sections must be designed accordingly.

Attenuators made up of a number of shunt and series arms are called ladder attenuators and may be built in a great variety of forms. Several of the commoner forms are shown in Fig. 63. The ladder attenuator of Fig. 63a is composed of a series of L sections of the type in Fig. 61d, with each designed to have an input resistance equal to the load resistance $R_L$ and an attenuation equal to the desired attenuation per switch step. The beginning of the series of attenuators is shunted by a resistance $R_s$ given by the relation[1]

[1] The resistance $R_s$ is called the iterative impedance of the attenuator and is the output resistance that would be obtained if there were an infinite number of sections to the ladder attenuator. Placing a resistance $R_s$ across the beginning of the attenuator is therefore equivalent to placing an infinite number of sections to the left of the beginning.

$$R_s = \tfrac{1}{2}(\sqrt{R_1(R_1 + 4R_2)} + R_1) \tag{35}$$

With this arrangement the resistance across *aa* is the same for all positions of the tap switch *S* and is the resistance formed by $R_s$ and $R_L$ in parallel, *i.e.*, $R_sR_L/(R_s + R_L)$. This is because the part of the attenuator to the left of every tap has a resistance equal to the iterative resistance $R_s$, while that to the right has a

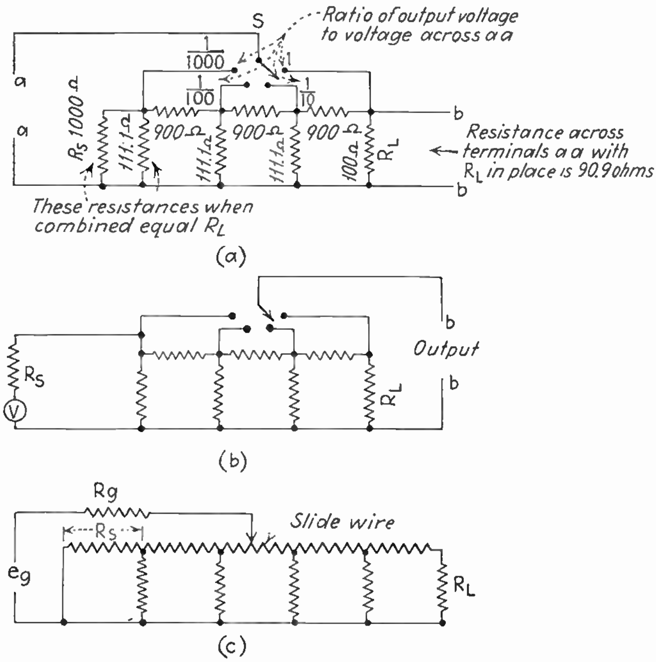



FIG. 63.—Typical ladder-type attenuators. The numerical values shown for *a* are where the attenuation $\alpha$ per section is 10, and where the load resistance $R_L$ is 100 ohms.

resistance equal to the load resistance $R_L$. If now a constant voltage $E$ is maintained between the terminals *aa*, the voltage across the load will be the potential $E$ reduced by the total attenuation of the L sections between the tap in use and the load, as given by Eq. (31*b*).

A variation of the foregoing ladder attenuator is to utilize resistance $R_L$ of Fig. 63*a* merely as a means of terminating the attenuator rather than as the load resistance. The voltage developed across $R_L$ is then considered to be the output voltage that is applied to the actual load that will be connected to *bb* in parallel with $R_L$. The output resistance of such an arrangement then has the constant value $R_L R_s/(R_L + R_s)$ for all settings of the switch, provided the current through *aa* is kept constant.

Numerical values for a specific case are shown in Fig. 63*a* and can be used to check the design procedure.

The ladder attenuator of Fig. 63*b* differs from that of Fig. 63*a* only in that the input is supplied by a voltage $V$ acting through a resistance equal to the iterative impedance of the attenuator as calculated by Eq. (32). The attenuation is now varied by switching the load connected across *bb*. In this arrangement the output resistance of the attenuator is the resistance formed by $R_s$ and $R_L$ in parallel and so is $R_s R_L/(R_s + R_L)$, and the relative attenuation at the different positions of the switch is determined only by the attenuator sections to the left of the tap in use.

A continuously adjustable ladder type of attenuator is shown in Fig. 63*c* and is the same as the step attenuator of Fig. 63*a* except that the tap switch has been replaced by a slider that permits continuous variation of the input point. As actually constructed, the resistance shown horizontally along the top is an ordinary slide-wire potentiometer to which shunt resistances have been connected at regular intervals to form a series of L sections. Such an attenuator is inexpensive to construct, and maintains its input and output resistances within narrow limits for all attenuations except extreme values. In particular, it will be noted that the input impedance can be made to have the correct value every time the slider passes over a point at which a shunt element is connected, while the output resistance will be substantially constant for all except very low attenuations, when the input connection is so close to the receiver that the shunting effect of the generator is appreciable.

The proper attenuator to use depends upon the requirements that must be met. The T attenuator is used where the insertion of the attenuator must not alter the impedance relations existing in the circuit. The L attenuator, on the other hand, can be used when the impedance need be kept constant on only one side of the attenuator. The continuously adjustable ladder attenuator will not keep its impedance constant on either side, but neither its input nor output resistance varies as much as the resistance on the variable side of an L attenuator. Ladder-type attenuators such as shown in Figs. 63*a* and 63*b* are used extensively for the production of known voltages or currents too small to be measured directly.

Attenuators are often employed in audio-frequency circuits at very low power levels, followed by large amounts of amplification. It is then essential that adjustment of the attenuator introduce negligible noise voltage. This is accomplished by protecting sliding contacts from dust, by using adequate pressure at the contact, and by making the two contacts of the same metal to avoid contact potentials.

*Decimal Attenuators.*[1]—A decimal attenuator is a system of attenuators so arranged that a voltage or current can be reduced

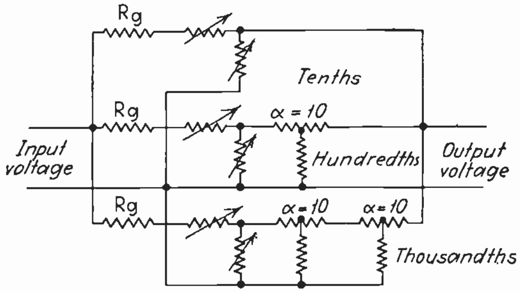



FIG. 64.—Circuit diagram of decimal attenuator.

in decimal fractions. Such an attenuator is shown in Fig. 64, where any voltage from 0.001 to 1.0 times the input voltage can be obtained in steps of 0.001 volt. This decimal attenuator includes three similar L-type attenuators designed to operate between equal generator and load resistances and to have a constant output resistance equal to the load resistance. Each attenuator is adjustable in steps such that $\alpha$ can be made 0, 0.1, 0.2, 0.3, etc., up to 1.0. The second of these L attenuators feeds into a single T attenuator having an input and output resistance equal to the output resistance of the L section, and having $\alpha = 10$, while the third of these attenuators delivers its output to two such T sections in tandem. These three systems hence give output voltages in steps of 0.1, 0.01, and 0.001, respectively. The outputs of all three attenuating systems are connected in parallel, so that when the output terminals are open-circuited

[1] Decimal attenuators as described here were developed by the General Radio Company.

the load resistance for any one attenuator system consists of the output resistance of the other two systems in parallel. The output voltages of the three attenuators are thus superimposed upon each other, and add up directly without being affected by the fact that the load resistance for any one attenuator is supplied by the other two attenuators which at the same time produce their own output voltages. The fact that there is a mis-match of resistance at the output need not be corrected for, since this changes all output voltages by the same percentage and so does not disturb the relative outputs. The output resistance to any load connected across the output terminals is the output resistances of the three attenuator systems in parallel. The addition of such a load merely increases the mis-match on the output side of the attenuator, and changes all output voltages by the same percentage without altering the relative values at the different attenuator settings.

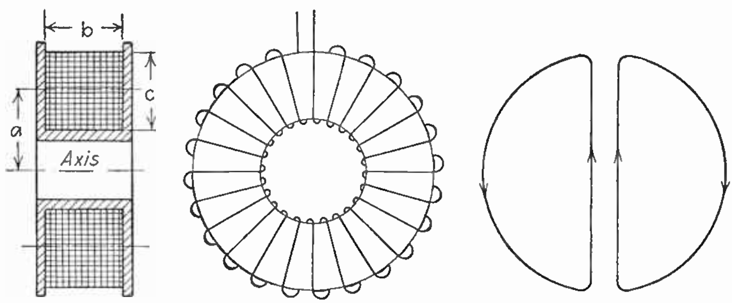
**24. Self- and Mutual-inductance Devices.**—The desirable properties of a standard inductance are mechanical stability, low temperature coefficient, a resistance which is low and varies as little as possible with frequency, a minimum of external magnetic field, and in the case of variable inductances a minimum of error from wear at the bearings.

Inductance standards for use at audio frequencies are normally air-cored coils wound with litz wire in order to minimize skin effect and make the alternating-current resistance substantially the same as the direct-current value throughout the audio-frequency range. Methods of arranging the winding are shown in Fig. 65. The conventional multilayer coil shown at *a* gives the largest inductance in proportion to the direct-current resistance, but it has the disadvantage of a large external field.[1] Inductances of this type will have minimum loss when proportioned so that $b = c$, and $c/a = 0.66$. The toroid has no external field but requires a very large amount of wire in proportion to the inductance obtained, and so has high direct-current resistance. The double-D winding is intermediate in its characteristics between the toroid and the multilayer coil, with respect to both external

[1] The design of standard inductances of this type is thoroughly discussed by H. B. Brooks, Design of Standards of Inductance, and the Proposed Use of Model Reactors in the Design of Air-core and Iron-core Reactors, Bur. of Std., *Jour. of Res.*, vol. 7, p. 289, August, 1931.

field and direct-current resistance. This type of coil has the advantage that the individual coils can be wound to approximately the correct number of turns, impregnated, taped, and mounted, all before final adjustment. The adjustment to give the exact value of inductance desired is then made by shifting the relative position of the two coils slightly with respect to each other before clamping them into final position.

Fixed inductances for use at radio frequencies should have low losses, good mechanical stability, and a minimum of dielectric loss. Litz wire is desirable for frequencies up to and including




(a) Multilayer Coil Rectangular Cross-section (b) Toroid (c) Double - D (coils connected in series)

FIG. 65.—Methods used to wind inductance standards for audio frequencies.

broadcast frequencies. The construction must be substantial and the form should be of good dielectric material with the coil protected from moisture by suitable "dope." In order to keep distributed capacity low, the coil should be a single-layer solenoid if possible; otherwise bank winding or similar arrangement must be employed to avoid excessive distributed capacity.

The temperature coefficient of an inductance coil depends primarily upon the construction. If the copper wire is allowed to expand and contract according to its own temperature coefficient, the inductance will vary approximately 17 parts in a million per degree centigrade. The winding form should if possible have a temperature coefficient of expansion that approximates this value in order that temperature cycles will not cause stretching of the copper and hence mechanical changes in its dimensions. No entirely suitable material appears to be available, although some grades of porcelain have the desired expansivity. Wood impregnated with beeswax appears to be about the most suitable material easily obtained, but is far from ideal. Bakelite, hard rubber, and fiber are not recommended because of their high coefficients of expansion.

An ingenious coil having zero temperature coefficient has been developed by Griffiths[1] and is illustrated in Fig. 66. Here the wire is wound upon sections of a cylinder that are held in position by a different material. The two materials are chosen so that, as the temperature increases, the rates of expansion across the diameter and along the length are just enough different to compensate for each other and keep the inductance constant. This result is possible because increasing the length of a coil reduces the inductance, while increasing the diameter increases

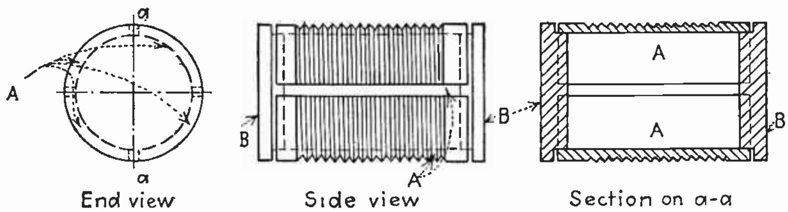



Fig. 66.—A coil having zero temperature coefficient of inductance. The material *A* controlling the expansion in an axial direction has a different coefficient of expansion from the material *B* controlling the coefficient of expansion in a radial direction, and the two can be so related that the resultant coefficient for inductance is zero.

the inductance. The latter effect is more important than the former if the expansion in both directions is the same. This principle can be applied to both multi-layer and single-layer coils.

*Variable Inductances.*—A variable inductance is essentially two fixed inductances connected in series and so arranged that the coefficient of coupling between them can be varied from a large positive value to a large negative value. In such an inductance it is essential that good mechanical stability be obtained and that the maximum coefficient of coupling be as near unity as possible.

The two types of variable inductances commonly employed are shown in Fig. 67. In the variometer a rotating coil is wound upon a spherical section and mounted inside a corresponding

[1] See W. H. F. Griffiths, Notes on Standard Inductances for Wavemeters and Other Radio Frequency Purposes, *Exp. Wireless and Wireless Engineer*, vol. 6, p. 543, October, 1929; W. H. F. Griffiths, Inductance for Radio Frequencies, *Wireless Engineer and Exp. Wireless*, vol. 11, p. 305, June, 1934.

fixed coil wound upon a section of spherical surface of slightly larger radius. The two coils are commonly adjusted so that their self-inductances are equal. In the Brooks inductometer the same result is achieved in a slightly different way. The advantage of the Brooks inductometer is that by proportioning the coils as shown in the figure the inductance varies linearly with the angle of rotation.[1] Both types of variable inductances shown in Fig. 67, when properly made, will have a maximum

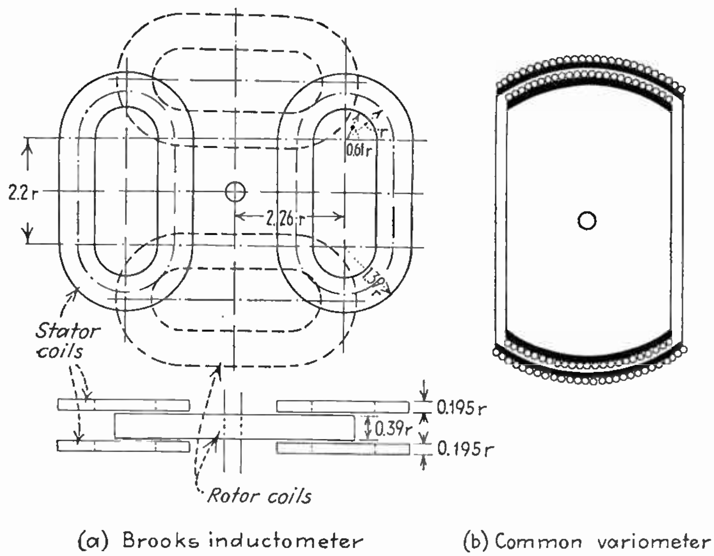



FIG. 67.—Types of variable inductances commonly employed as standards of self- and mutual inductance.

coefficient of coupling such that the ratio of maximum to minimum inductance will exceed 10 to 1.

*Mutual-inductance Standards.*—Any variable inductance can be converted into a variable mutual inductance by bringing out the fixed and moving coils to separate binding posts. This is ordinarily done in all variable inductance standards, and permits such standards to be used as variable mutual inductances by the removal of a jumper.

*Decade Inductances.*—Inductances can be built up into decade units in the same way as resistances. One way of accomplishing

[1] For detailed design information on the Brooks inductometer see A Variable Self and Mutual Inductor, Bur. Std., *Sci. Paper* 290.

this is to cover an entire decade with a single tapped coil. It is difficult, however, to locate the taps properly unless the coil is wound by hand. A more satisfactory arrangement is to build each decade up from four untapped inductances controlled by a gang switch as shown in Fig. 68. Care must be taken in mounting the coils to insure that they have substantially zero coupling; and if compactness is at all important, it is necessary to use toroidal coils.

Coils wound upon permalloy or iron-dust rings are particularly suitable for audio-frequency laboratory inductances up to about 1 henry. The dust rings are sufficiently stable in their characteristics to maintain the inductances accurate to closer than 1 per cent under ordinary conditions provided that excessive direct-

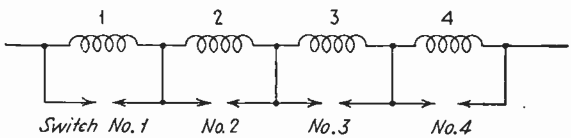



FIG. 68.—Switching arrangement for decade inductance. A four-gang switch is required and is wired so that the undesired inductances are short-circuited.

current magnetizations are not allowed. Such coils have very low losses at audio frequencies (*i.e.*, $Q$ in the order of 25 to 100), and do not vary their inductance appreciably with reasonable flux densities. The core losses in such inductances increase as the frequency is raised, but since the losses are small this is not particularly important for most applications.

**25. Capacity Devices.**—The characteristics desirable in standards of capacity are permanence of calibration, low temperature coefficient, low losses, and losses that vary according to a known law.

Standards of capacity covering the range up to about 0.002 $\mu$f are always variable condensers. The construction must be rugged, good bearings free of backlash are essential, and rotor stops are to be avoided because they give rise to mechanical shocks. A slow-motion device is desirable to permit accurate adjustment of the capacity. The entire condenser must be mounted in a metal-lined container connected to the rotor plates so that neighboring objects will have no effect on the capacity.

The temperature coefficient of capacity of a variable condenser is proportional to $\alpha_1^2/\alpha_2$ where $\alpha_1$ and $\alpha_2$ are the coefficients of expansion of the material of the rotor plates and of the spacing washers, respectively. Metals suitable for use in condensers have coefficients of expansion of the order of 15 parts per million per degree centigrade, so ordinary condensers can be expected to have a temperature coefficient of capacity of the same order of magnitude.[1]

The best standard variable condensers have the dielectric arranged so that the electrostatic stress in the insulation is not affected by the capacity setting of the condenser. Inasmuch as nearly all the losses of a good condenser occur in the material that insulates the rotor and stator plates from each other, a condenser constructed in this way will have a power loss that is substantially independent of capacity until the frequency becomes so high (above perhaps 5 mc) that eddy-current losses in the plates can no longer be neglected. Such a condenser can therefore be represented by the circuit of Fig. 69*a*, in which $C_1$ is the loss-free part of the capacity, and includes the variable portion, while $C_0$ represents the capacity associated with the solid dielectric and having the loss resistance $R_0$. It is apparent that the solid dielectric should have the lowest possible power factor (hard rubber and isolantite are preferred), and should be arranged in such a way as to contribute as little as possible to the condenser capacity.

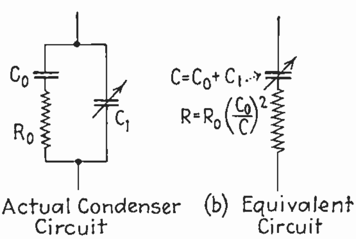


FIG. 69.—Actual and equivalent circuits of condenser with losses.

*Fixed Condensers.*—Standards of capacity larger than about 0.002 $\mu$f are normally fixed condensers with mica dielectric. When properly constructed, such condensers have very low losses and are stable in their characteristics. The power factors that can

[1] A modified form of variable condenser has been developed which can be made to have zero temperature coefficient of capacity. See W. H. F. Griffiths, Further Notes on the Calibration Permanence and Overall Accuracy of the Series-gap Precision Variable Air Condenser, *Exp. Wireless and Wireless Engineer*, vol. 6, p. 23, January, 1929.

be expected depend upon the construction, capacity, and frequency, and range from about 0.05 to 0.25 per cent for all frequencies except those so high that eddy-current losses in the plates and leads become appreciable. The temperature coefficient of capacity depends upon the details of construction but will normally be in the range from 50 to 100 parts per million per degree centigrade.

It is expensive to adjust fixed condenser standards exactly to a predetermined capacity, because unless special precautions are taken the final capacity may change during impregnation and mounting, and it is then too late to make adjustments. As a result, most standards of fixed capacity will depart from exact decimal values by appreciable amounts (*i.e.*, as much as 1 per cent) although any condenser can have its capacity measured and hence known to a very high degree of precision.

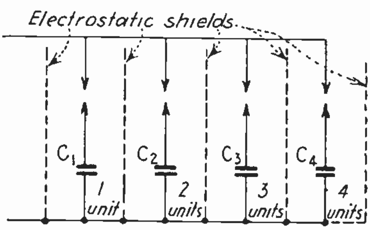



FIG. 70.—Switching arrangement for decade condenser. This requires a four-gang 11-position switch so wired up that the necessary condensers are connected across the circuit for the various switch positions. The electrostatic shields shown are necessary if the upper sides of the condensers and the associated wiring are not spaced so as to have negligible direct capacities to introduce errors.

*Decade Condensers.*—In grouping condensers together to form decade units, it is desirable to employ the smallest possible number of condensers because of the cost of accurately adjusted units. A suitable arrangement is shown in Fig. 70 and permits a full decade to be covered with four condensers controlled by a four-gang 11-position switch. Decade condensers suitable for general laboratory use may be assembled from small mica radio condensers having capacities slightly less than the desired values, which are then padded out to the desired value with adjustable "trimmer" condensers of the type employed to line up intermediate-frequency transformers in radio receivers. This method of construction is suitable for decade units having steps of 0.001 and 0.01 $\mu$f. A variable condenser having a maximum capacity in excess of 0.001 $\mu$f (such as a four-gang radio condenser with all sections in parallel) should be used for interpolating between 0.001-$\mu$f steps. Mica condensers suitable for 0.1-$\mu$f steps are very expensive, and unless low losses are very essential it is usually desirable to use a good-grade paper condenser.[1]

Decade condenser boxes should be mounted in a metal container so that their capacitance will not be affected by outside objects. It is also necessary to shield each individual condenser and its associated switch if the capacities of the various steps are to add up properly. Suitable shielding is shown in Fig. 70. If the shielding is omitted, direct capacities will exist between the upper sides of the various condensers, as illustrated in Fig. 71, and this introduces errors. Thus when one of the condensers $C_1$ is switched into the circuit, the capacity added is not $C_1$, but rather $C_1$ plus shunting capacities such as the capacity consisting of $C_2$ in series with $C_{12}$. On the other hand, when all capacities are switched in, the total capacity is $C_1 + C_2 + C_3 + C_4$, and the direct capacities $C_{12}$, $C_{23}$, etc., have no effect. The shield replaces the direct capacities by capacities to ground which are in parallel with the individual unit and can be allowed for when the unit is constructed and adjusted.

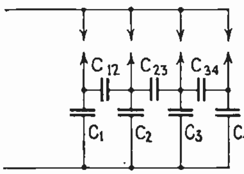


FIG. 71.—Unshielded decade condenser, illustrating direct capacities that introduce errors.

The effects of unshielded direct capacities are particularly important in the 0.001- and 0.01-$\mu$f decades, since the error introduced by the direct capacity is not negligible in comparison with the size of these steps. In the case of larger decades the error is normally negligible when expressed on a percentage basis.

All decade condensers have a residual capacity at zero setting. This should be kept as low as possible by using the simplest switches but can never be reduced to zero. The minimum capacity of a decade condenser should, therefore, be determined and marked on the case. The condenser is used with the knowledge that this capacity is always present and that the decade switches merely introduce known additional amounts of capacity.

[1] There are a series of telephone (Western Electric) condensers peculiarly well suited to building up decade units having steps of 0.01 $\mu$f and larger. These consist of a large number of individual condensers of various sizes potted in a single can, and are relatively inexpensive. It is possible to build up any desired capacity to a high degree of accuracy by merely combining units in the one can.

## CHAPTER V

# MEASUREMENT OF FREQUENCY

**26. Standards of Frequency.**—The fundamental standard of frequency is the period of rotation of the earth. This can be measured with great accuracy by astronomical methods, and might be thought of as a standard frequency of one cycle per day. All standards of frequency must ultimately be referred to this fundamental source of frequency for calibration purposes.

Practical frequency standards can be classified as either primary or secondary standards. A primary standard of frequency is an oscillator which generates a frequency that is very constant over long periods of time, and which is checked against the earth's rotation at regular intervals. Primary standards are commonly provided with harmonic and subharmonic generators so as to give a large number of different frequencies all related to, and just as accurate as, the oscillation originally generated. Secondary standards of frequency are stable oscillators which have their frequency checked periodically against a primary standard, and are also often provided with harmonic and subharmonic generators.

Where the high accuracy of primary and secondary standards is not required, frequency can be measured by wavemeters, heterodyne frequency meters, or a Lecher wire system. A wavemeter is an ordinary tuned circuit provided with a calibration giving resonant frequency as a function of the setting of the tuning condenser. A heterodyne frequency meter is an oscillator provided with a similar frequency calibration. The Lecher wire arrangement is suitable for measuring very high frequencies, and consists of a resonant transmission line upon which the wave length is measured directly by determining the distance between adjacent nodes.

**27. Primary Standards of Frequency.**—All commercial primary standards of frequency used in radio work employ a carefully designed crystal oscillator.[1] The crystal is operated in a constant-temperature oven,[2] is preferably cut so that it has negligible temperature coefficient,[3] and every other precaution that will improve the frequency stability is taken.

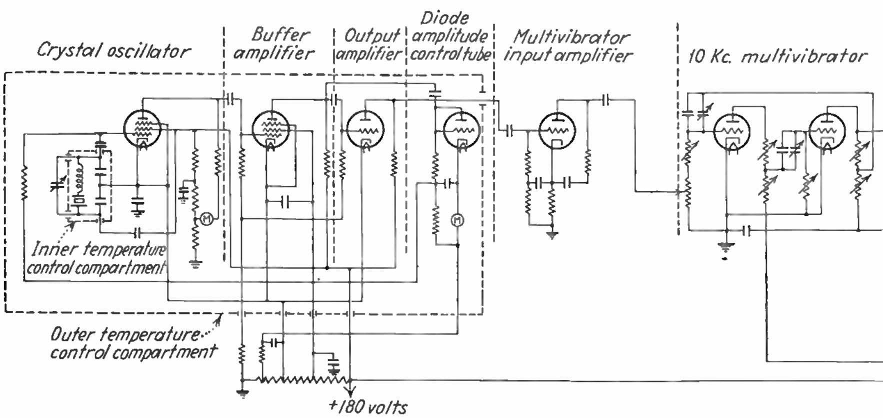



FIG. 72.—Schematic circuit diagram of

The circuit diagram of a commercial primary frequency standard is shown in Fig. 72.[4] A modified form of Colpitts oscillator circuit is used in an electron-coupled arrangement.

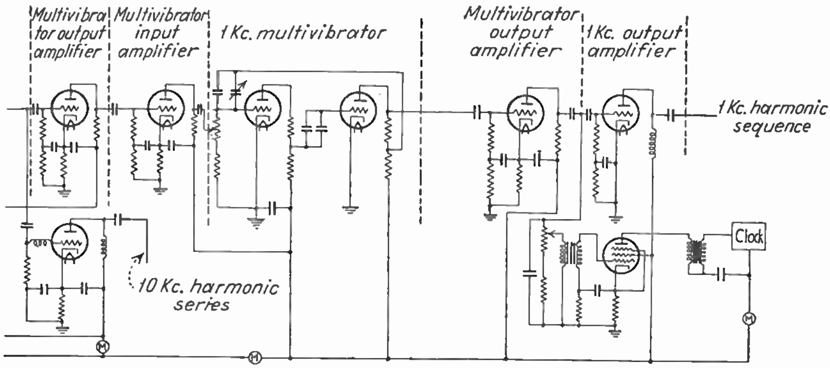
[1] Other possible primary standards are precision clocks, which have satisfactory accuracy but develop a frequency too low (*i.e.*, one cycle per second) to be easily used in measuring radio frequencies, and the tuning fork. Electrically driven forks have been used for primary standards, and probably have about the same ultimate possibilities as the crystal oscillator, although the latter has been more fully developed. For further information on precision forks see J. W. Horton, N. H. Ricker, and W. A. Marrison, Frequency Measurement in Electrical Communication, *Trans. A.I.E.E.*, vol. 42, p. 730, 1923; E. Norrman, A Precision Tuning Fork Frequency Standard, *Proc. I.R.E.*, vol. 20, p. 1715, November, 1932.

[2] The principles involved in such ovens are described in detail by J. K. Clapp, Temperature Control for Frequency Standards, *Proc. I.R.E.*, vol. 18, p. 2003, December, 1930. Also see W. A. Marrison, Thermostat Design for Frequency Standards, *Proc. I.R.E.*, vol. 16, p. 976, July, 1928.

[3] For methods of accomplishing this see F. R. Lack, Observations on Modes of Vibrations and Temperature Coefficients of Quartz Crystal Plates, *Proc. I.R.E.*, vol. 17, p. 1123, July, 1929; F. R. Lack, G. W. Willard, and I. E. Fair, Some Improvements in Quartz Crystal Circuit Elements, *Bell System Tech. Jour.*, vol. 13, p. 453, July, 1934.

[4] This figure is based upon information supplied to the author by J. K. Clapp of the General Radio Company.

The reactances associated with the crystal are so proportioned that the crystal operates almost exactly at its series resonant frequency. Automatic amplitude control which has both "amplified" and "delay" action is employed to keep the ampli-




commercial primary frequency standard.

tude of oscillations constant with changes in tube constants and electrode voltages. The crystal operates at 50 kc and consists of a bar rigidly clamped at its mid-point (which is a nodal point for these vibrations), with the electrodes silvered directly upon the crystal. Detrimental effects from resonances in the air are eliminated by baffle plates mounted one-fourth wave length from the ends of the crystal. The entire crystal oscillator with its buffer amplifiers and amplitude-control tube is mounted in a constant-temperature oven, and the crystal itself is mounted in an inner compartment having an additional thermostat and heater. The output of the crystal oscillator unit controls a 10-kc multivibrator, which in turn controls a 1-kc multivibrator that is used to drive a synchronous clock that keeps correct time when supplied with exactly 1000 cycles. Provision is made for accurately comparing the time kept by the clock with time signals sent out by government radio stations and obtained by them from observatories. Small adjustments in crystal frequency as found necessary can be made with the variable condenser shunting the crystal holder.

Harmonic sequences for measuring purposes are obtained by additional output amplifiers as desired. These usually include a 10-kc sequence, and in some cases 1-kc and 50-kc sequences. In the output amplifier tubes, the plate load impedance consists of a small inductance such that the high-order harmonics (those above the one hundredth) will be accentuated in comparison with the low-order harmonics. Since a single multivibrator will develop harmonics up to the three hundredth to five hundredth that are detectable on a radio receiver, it is possible to obtain a great number of frequencies, all of which have exactly the same degree of precision as the frequency of the crystal oscillator.

All commercial primary standards of frequency will maintain their frequency constant to within one part in a million over long periods of time. By checking daily against time signals, the degree of precision is considerably higher and can be expected to be nearer one part in ten million. The accuracy which this represents is illustrated by the fact that one ten-millionth of the distance between New York and San Francisco is approximately 16 in.

**28. Secondary Standards of Frequency.**—Secondary frequency standards may be of a variety of types depending upon the requirements of the individual case. For general measurement work the most accurate secondary frequency standard is a good crystal oscillator combined with a single multivibrator for generating a series of harmonics. This is essentially a simplified primary frequency standard with the omission of the synchronous clock and of some of the refinements that add greatly to cost and relatively little to stability. A typical circuit diagram is shown in Fig. 73. A properly designed crystal oscillator operating in the frequency range 30 to 100 kc and provided with a simple constant-temperature oven will maintain its frequency with an accuracy of at least 20 parts in a million almost indefinitely and can be depended upon to give considerably greater accuracy if periodically checked against a primary standard.

*Frequency Monitors.*—Government regulations require that radio transmitting stations, particularly broadcast stations, maintain their frequency very close to the assigned value. The method ordinarily employed at the transmitter to insure that these regulations are being complied with is to provide a simplified secondary frequency standard against which the transmitted frequency is continuously compared. This equipment is commonly called a station frequency monitor, and consists of a suitable crystal oscillator with some simple means to give a continuous indication of the transmitted frequency in terms of the monitor. One type of frequency-monitoring equipment widely used by broadcast stations consists of a high-grade crystal oscillator employing a crystal which generates a frequency that is exactly 1000 cycles different from the assigned frequency. Outputs derived from the radio transmitter and the frequency monitor are simultaneously applied to a detector that will hence produce a beat note of exactly 1000 cycles when the transmitter

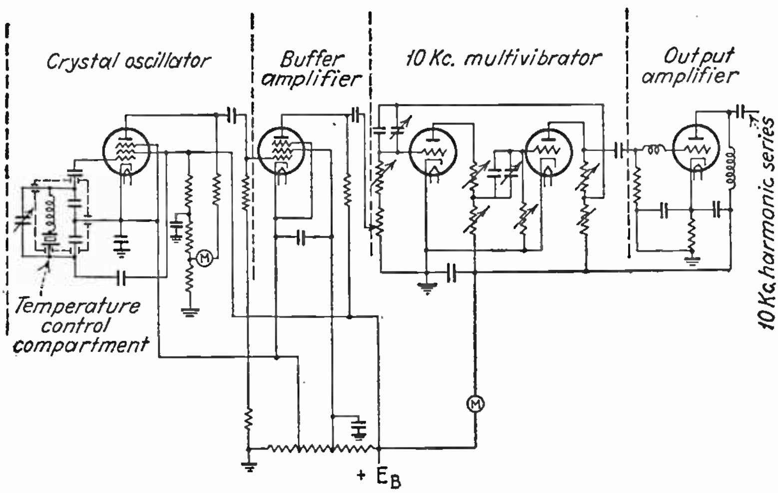



FIG. 73.—Schematic circuit diagram of secondary frequency standard.

has the correct frequency. This beat note is supplied to a frequency meter which covers the range 950 to 1050 cycles and so indicates continuously the frequency of the transmitter with respect to the monitor. A modification of this monitoring system is to grind the crystal of the monitor to the exact frequency assigned to the transmitter, and then to adjust the transmitted frequency to within one or two cycles per second of the monitor frequency. In this case the difference frequency is indicated by beats of a direct-current meter in the plate circuit of a detector to which both signal and monitor frequencies have been applied.

**29. Wavemeters and Heterodyne Frequency Meters.**—A wavemeter is a resonant circuit tuned by an ordinary variable condenser and provided with a calibration that gives the resonant frequency in terms of the condenser setting. The coil and condenser of the wavemeter should be stable with respect to age and handling, and the temperature coefficient should be as small as possible. Where the maximum possible precision is desired, a thermometer should be mounted on the wavemeter, and correction made for temperature. It is also very desirable that the tuned circuit have low losses since the accuracy with which frequency can be measured is determined primarily by the $Q$ of the tuned circuit.

Resonance between the wavemeter circuit and the oscillations being measured can be determined by a low-resistance thermocouple milliammeter directly in series with the tuned circuit or coupled to it by a small coil rigidly mounted in fixed relation to the wavemeter inductance. A neon lamp across the tuning condenser is also sometimes used in making rough measurements.

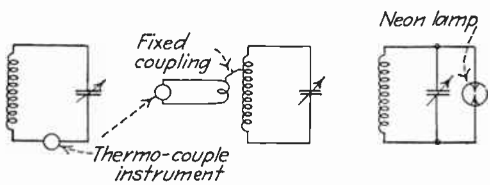



FIG. 74.—Methods of indicating when a wavemeter is tuned to resonance with the frequency being measured.

These arrangements are shown in Fig. 74. Resonance between a wavemeter and a small oscillator can also be detected by observing the effect of the wavemeter upon the d-c plate current of the oscillator, which always jumps as the wavemeter is brought into resonance, or the effect on the frequency, which will shift slightly in the same situation. If the oscillator is of low power and the coupling to the wavemeter sufficiently close, the oscillations will cease as the wavemeter is brought into resonance. This phenomenon is often employed in rough measurements but is not very accurate because of the close coupling required. In order to obtain accuracy it is necessary that the coupling between the wavemeter and neighboring objects be very loose, for otherwise the impedance which is coupled into the wavemeter circuit will alter the calibration.

The accuracy with which frequency can be determined by a wavemeter ranges from 1.0 to 0.1 per cent, with the value in any specific case depending upon the construction of the instrument, the $Q$ of the tuned circuit, and whether or not correction is made for temperature. In general the accuracy with which one can determine a frequency is less than the accuracy with which the dial setting may be read because of the fact that the flat top of the resonance curve makes it impossible to tell exactly when the proper setting has been obtained.

A heterodyne frequency meter is an oscillator for which a calibration has been made of frequency as a function of tuning capacity. The oscillator should be of a type that has good frequency stability. The most satisfactory oscillator is the electron-coupled type (see Sec. 72) since this has excellent frequency stability and its output is rich in harmonics. Dynatron oscillators are also frequently used, and have frequency stability as good as or better than the electron-coupled oscillator if the grid bias is adjusted so that oscillations just barely start. A dynatron oscillator operated in this way does not produce harmonics, however. The tuned circuit of the oscillator must be completely shielded so that neighboring objects will not affect the frequency, and the output must be so arranged that there is no reaction on the frequency.

A well-made heterodyne frequency meter will maintain its calibration to better than 0.1 per cent over long periods of time provided correction is made for temperature changes. Since the principal factor other than temperature that affects the frequency is aging and tube changes, the precision over short periods of time, *i.e.*, a few hours to a few days, is much higher and may even reach 10 to 20 parts in a million.

**30. Methods of Comparing Frequencies.**—When the standard develops a series of fixed frequencies, as is the case with all primary and secondary frequency standards, the measuring problem is one of comparing the unknown frequency with known frequencies which are slightly different. There are two principal methods by which this can be accomplished: These are, *first*, the direct measurement of the difference between the unknown and the nearest known frequency,[1] and, *second*, direct interpola-

[1] A modification is to employ several successive heterodynings between the unknown and the known frequencies to reduce the order of magnitude of the difference to a small residual that can be measured to a small fraction of a cycle. This method requires a rather elaborate equipment, however, and so is not extensively used. It is described by F. A. Polkinghorn and A. A. Roetken, A Device for the Precise Measurement of High Frequencies, *Proc. I.R.E.*, vol. 19, p. 937, July, 1931.

tion between the nearest known frequencies by means of an interpolation oscillator.

The direct measurement of the difference frequency is the more accurate but requires somewhat more accessory equipment than does the interpolation method. The former method is normally employed when comparing against a primary frequency standard, while the latter finds its chief usefulness in conjunction with secondary frequency standards.

An understanding of the problems involved in comparing frequencies can best be gained by considering the detailed measuring procedure required. Take first the problem of determining the frequency of a radio signal by the difference-frequency method. The principal steps are as follows:[1]

1. The signal of frequency $f_x$ is tuned in on a radio receiver which is preferably of the oscillating-detector type.

2. A heterodyne frequency meter is then adjusted to zero beat with the signal. This can be accomplished to closer than one cycle per second by adjusting the oscillating detector of the radio receiver to give a beat note of approximately 1000 cycles with the signal. As the heterodyne frequency meter is then brought into zero beat with the signal, the 1000-cycle beat note will wax and wane at a rate that corresponds to the difference between the frequencies of the heterodyne frequency meter and the signal. This waxing and waning should be independent of the frequency of the oscillating detector; and if a slight change in the adjustment of the oscillating detector alters the rate, then the heterodyne frequency meter has been set to zero beat with the oscillating detector and not with the signal. This procedure effectively transfers the signal to the heterodyne frequency meter and eliminates such factors as fading and failure of signal. It also gives the approximate frequency of the signal from the calibration of the heterodyne frequency meter.

3. The output of the heterodyne frequency meter is then loosely coupled to a non-oscillating detector and the 10-kc harmonic sequence of the standard frequency generator is introduced into the circuits of the detector. The non-oscillating detector can, if desired, be the original radio receiver with the regeneration control set back and the antenna disconnected. The detector output will contain an audio-frequency beat note of less than 5000 cycles as a result of heterodyne action between the signal frequency and the nearest harmonic of the 10-kc sequence.

[1] This procedure is essentially that given in the bulletin, Frequency Measurements at Radio Frequencies, issued by the General Radio Company. Further information on the general problem of comparing frequencies is to be found in this bulletin, and in the paper by J. K. Clapp, Interpolation Methods for Use with Harmonic Frequency Standards, *Proc. I.R.E.*, vol. 18, p. 1575, September, 1930.

4. The audio-frequency beat note is now measured by an audio-frequency bridge or by comparison with a calibrated audio-frequency oscillator.
5. The unknown frequency is then the frequency of the nearest harmonic of the 10-kc sequence plus or minus the audio-frequency beat note. It is possible to tell whether the audio note must be added or subtracted by increasing the frequency of the heterodyne frequency meter slightly. If this increases the audio frequency, one must add the difference frequency to the harmonic just lower than the unknown signal frequency, while a decrease in the audio note indicates that subtraction from the harmonic just higher than the signal frequency is required.

The foregoing procedure must be modified at very high and very low radio frequencies because of practical considerations. At frequencies above about 3500 kc the harmonics of the 10-kc multivibrator are not always of sufficient amplitude to be detected, while at very low frequencies, *i.e.*, below about 100 kc, the difference between unknown signal frequency and the nearest harmonic becomes such a large percentage of the unknown frequency as to introduce an appreciable percentage error. There is, furthermore, the practical inconvenience of providing the heterodyne frequency meter with sufficient coils to cover all frequency ranges. These limitations are readily overcome by the expedient of using suitable harmonics. Thus in measuring very high frequencies one may set a harmonic of the heterodyne frequency meter to zero beat with the signal and then measure the fundamental frequency of the heterodyne frequency meter by the procedure outlined above. In the case of low signal frequencies, on the other hand, one may set the fundamental of the heterodyne frequency meter to zero beat with a harmonic of the unknown frequency, and then measure the fundamental frequency of the heterodyne frequency meter.

The accuracy with which frequencies can be measured by the foregoing methods depends primarily upon the accuracy of the frequency standard, the precision with which the audio-frequency beat note can be measured, and the stability of the heterodyne frequency meter over the time required to complete the measurement. With suitable equipment the errors exclusive of the error in the standard frequency should not exceed a few cycles per second.

In order to clarify the procedure, some typical examples will be now considered. Assume first that it is desired to measure a frequency which is supposed to be 1693 kc. Following the procedure that has been outlined, we should find from the calibration of the heterodyne frequency meter that after Step 2 is completed the signal would lie between the 169th and the 170th harmonic of the 10-kc sequence.[1] The audio beat note heard in Step 3 will be 3000 cycles and will be added to 1690 kc because an increase in the frequency of the heterodyne frequency meter makes the beat note increase. If a decrease in the audio note had been observed, the 3000 cycles would have been subtracted from the next higher harmonic, *i.e.*, subtracted from 1700 kc.

Assume next that the frequency to be measured is considerably higher, say 20,641 kc. It is now necessary to modify the procedure since detectable harmonics of the 10-kc sequence do not extend to such high frequencies. The procedure is to set a harmonic of the heterodyne frequency meter to zero beat with the signal. Thus if the tenth harmonic is used, the heterodyne frequency meter will be adjusted to 2064.1 kc.[2] The frequency 2064.1 is now measured by the identical procedure described above and the results multiplied by 10 to give the desired frequency.

Finally consider the problem of measuring a frequency supposed to be 31.6 kc, assuming that the heterodyne frequency meter does not operate below 200 kc. The 31.6-kc frequency is picked up on an oscillating receiver, which is made to oscillate so vigorously that harmonics of the signal frequency are gener-

[1] When very high-order harmonics such as the 451st and 452d, for example, are involved, the percentage difference between adjacent harmonics is so small that there may sometimes be doubt as to just which harmonics are being used. In such circumstances the simplest procedure is to provide certain harmonics with a distinguishing feature. Thus when the measuring is done with a 10-kc harmonic sequence, one can provide an auxiliary 100-kc multivibrator controlled from the same source. By modulating the output amplifier of this multivibrator with a distinguishing tone, such as 60 cycles, it is possible to locate every tenth harmonic of the 10-kc sequence with certainty, and intermediate harmonics may then be determined by counting from the nearest harmonic so marked. Thus in the case of uncertainty mentioned above, one would locate the 45th harmonic of the 100-kc multivibrator, which can be readily done without uncertainty. The harmonic in question would then be the 451st or the 452d depending upon whether it was the first or second harmonic of the 10-kc sequence beyond the harmonic marked with the 60-cycle distinguishing tone.

[2] The particular harmonic of the heterodyne frequency meter that is employed can be readily determined from a rough frequency calibration of the radio receiver, and the calibration of the heterodyne frequency meter.

ated. The fundamental of the heterodyne frequency meter is then set to zero beat with a convenient harmonic, say the eighth, of the signal. The resulting frequency of 252.8 kc is then measured by the procedure outlined above, and the results divided by 8 to give the unknown frequency.

*Direct-interpolation Method of Comparing Frequencies.*—The procedure for comparing frequencies by the direct-interpolation method is normally as follows: The signal is tuned in upon a radio receiver and is transferred to a heterodyne frequency meter as in the previous case. The output of the heterodyne frequency meter and the 10-kc harmonic sequence are then simultaneously applied to a detector, and the heterodyne frequency meter is adjusted to zero beat with the harmonics of the 10-kc sequence on either side of the signal frequency.[1] A linear interpolation between the three dial settings of the heterodyne frequency meter is then made to determine the unknown frequency.

The accuracy of this method depends largely upon the linearity of the relation between frequency and dial setting given by the heterodyne frequency meter, and upon the openness of the frequency calibration curve. The absolute frequency calibration of the heterodyne frequency meter does not enter into the result. The accuracy obtainable by the interpolation method is never so great as by the direct measurement of the frequency difference, but it can be made sufficient to be entirely satisfactory when secondary frequency standards are employed. The accuracy of the interpolation can be greatly increased by designing the interpolation oscillator to have an open scale, *i.e.*, a small number of cycles per division.

Some specific examples will make the interpolation details clear. It will be assumed that the same frequencies are to be measured as in the previous examples, that one has available a 10-kc harmonic sequence, and that the heterodyne frequency meter will cover the range 200 to 2000 kc. To measure the 1693-kc signal, one would tune this in on the radio receiver and adjust the heterodyne frequency meter to zero beat. The

[1] The zero-beat setting can be done very accurately by allowing the detector to oscillate at a frequency that gives a beat note of about 1000 cycles with the harmonic in question and observing the waxing and waning of the beats with the heterodyne frequency meter as described in Step 2 above.

calibration of the frequency meter would indicate that the signal was between the 169th and the 170th harmonic. Then

$$\text{Unknown frequency} = 10{,}000\left(169 + \frac{\theta_x - \theta_1}{\theta_2 - \theta_1}\right)$$

where $\theta_x$, $\theta_1$, and $\theta_2$ are respectively the settings of the heterodyne frequency meter corresponding to zero beat with the unknown frequency, the harmonic just below $\theta_x$ (169th), and the harmonic just above $\theta_x$ (170th). In this particular case one would find $(\theta_x - \theta_1)/(\theta_2 - \theta_1)$ would equal 0.3.

Consider next the problem of measuring a signal having a frequency of 20,641 kc. The twelfth harmonic of the heterodyne frequency meter could be set to zero beat with this signal, thus causing the heterodyne frequency meter to have a fundametal of 1720.08 kc. The calibration of the frequency meter would show that the frequency was very close to 1720 kc. The meter setting corresponding to zero beat with the 172d and 173d harmonics is then determined and interpolation made according to the following formula:

$$\text{Unknown frequency} = 12 \times 10{,}000\left(172 + \frac{\theta_x - \theta_1}{\theta_2 - \theta_1}\right)$$

where the notation follows the same system as used above. Here $(\theta_x - \theta_1)/(\theta_2 - \theta_1)$ would be found equal to 0.008.

Finally consider the problem of measuring a low frequency such as 31.6 kc. In this case the oscillating receiver would be overloaded until a harmonic such as the seventh was developed. The fundamental of the heterodyne frequency meter would be set to 221.2 to give zero beat with this harmonic. The calibration would indicate that the adjacent harmonics of 10 kc were the twenty-second and twenty-third, and interpolation between these by adjusting to zero beat with these harmonics and the unknown frequency would give, using the previous notation,

$$\text{Unknown frequency} = \frac{10{,}000\left(22 + \dfrac{\theta_x - \theta_1}{\theta_2 - \theta_1}\right)}{7}$$

In this case $(\theta_x - \theta_1)/(\theta_2 - \theta_1)$ would be found to be 0.12.

*Frequency Measurements with Heterodyne Frequency Meter.*—A heterodyne frequency meter is essentially a calibrated oscillator and is one of the most useful of all frequency-measuring devices. Its use in connection with accurate measurements involving primary and secondary standards has already been discussed, but for many purposes the accuracy of a heterodyne frequency meter, which is in the order of 0.1 per cent over long periods of time, is sufficient, without the use of any other standard.

In making frequency measurements with a heterodyne frequency meter the most satisfactory method is to tune the unknown frequency in on a radio receiver having an oscillating detector, and then to adjust the heterodyne frequency meter to zero beat with the signal, using the method described in Step 2 of the procedure outlined at the beginning of this section. In cases where the unknown frequency is outside the range covered by the heterodyne frequency meter, harmonic methods can be employed. Thus in measuring high frequencies, a harmonic of the heterodyne frequency meter may be adjusted to zero beat with the signal, while with low unknown frequencies a harmonic of the unknown can be measured. The particular harmonic employed in any case can be readily determined with an ordinary wavemeter or from a rough calibration of the radio receiver.

It is possible to equip a heterodyne frequency meter with a detector and audio amplifier in order to indicate beat notes between the oscillations of the frequency meter and a voltage of unknown frequency coupled into the circuits of the frequency meter. This arrangement, while convenient, will not give the maximum possible accuracy since the heterodyne frequency meter tends to synchronize with the injected frequency, and is thus drawn off frequency by an amount which depends upon the amount of voltage introduced.

**31. Radio Signals as Frequency Standards.**—Certain classes of radio signals are very useful as standards of frequency. Thus the Bureau of Standards carries on a regular schedule of transmissions at 5000 kc which have an accuracy which is the same as the primary frequency standard maintained by the bureau (*i.e.*, better than one part in one million). These signals can be used to check the accuracy of secondary standards and to calibrate heterodyne wavemeters and so make a primary standard unnecessary under most circumstances.

The simplest method of using the Bureau of Standards' 5000-kc transmission is to provide a 50-kc oscillator (either crystal or high-stability tuned-circuit type) which can have its fundamental frequency varied over a very narrow range.

The oscillator should be provided with a 50- and a 10-kc multivibrator. Inasmuch as these two multivibrators need not be used simultaneously, it is permissible to use a single multivibrator with a switch to change the circuit constants to give either 10- or 50-kc oscillations. A schematic diagram of the equipment required is shown in Fig. 75. The Bureau of Standard signals are tuned in on the oscillating-detector receiver and harmonics from the 50-kc multivibrator introduced into the receiver. The frequency of the 50-kc oscillator is then varied until the one-hundredth harmonic is exactly synchronized with the signals. By now switching to the 10-kc multivibrator one obtains a 10-kc harmonic sequence which at the moment has an accuracy equal to the Bureau of Standards' primary frequency standard. The accuracy after some time has elapsed will, of course, depend upon the stability of the 50-kc oscillator.

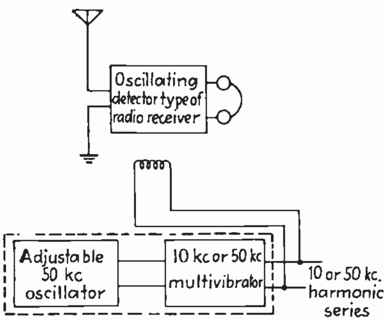



FIG. 75.—Schematic diagram of equipment for utilizing the Bureau of Standards 5000-kc standard-frequency transmissions. The multivibrator is provided with a switch that will change its fundamental frequency from 10 to 50 kc.

*The Use of Broadcast Stations as Frequency Standards.*—Broadcast stations are generally good sources of standard frequencies since they are in operation continuously, and are required to maintain frequency to within 50 cycles of the assigned value. The high-power stations do much better than this, normally keeping their carrier to within well under 10 cycles of the assigned value, which represents a precision about equal to that of a good secondary frequency standard, *i.e.*, about 10 to 20 parts in a million.

The most satisfactory method of using broadcast signals is to provide a 10-kc oscillator adjustable over a very narrow frequency range and having high stability, together with a controlled 10-kc multivibrator, as shown in Fig. 76. The desired station is tuned in on an ordinary broadcast receiver which does not have automatic volume control. The 10-kc harmonic sequence from the multivibrator is then applied to the receiver and the 10-kc oscillator adjusted until zero beat is obtained between the appropriate multivibrator harmonic and the broadcast signal. This zero-beat condition can be determined from the vibration of a sensitive direct-current meter in the detector plate circuit, or from the flutter of the received signal. The 10-kc harmonic sequence of the multivibrator is now as accurate as the frequency of the broadcast station, and can be used for measurement purposes.

Since all broadcast stations are assigned frequencies that are multiples of 10 kc, one can check the accuracy of the 10-kc

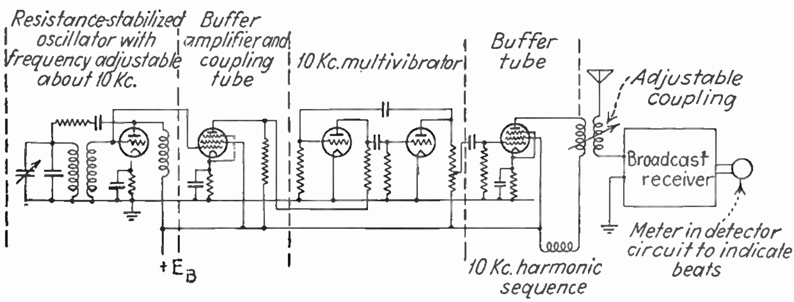



FIG. 76.—Circuit diagram of apparatus for producing a 10-kc harmonic sequence that is adjusted to high accuracy with the aid of carrier waves of broadcast stations.

harmonic sequence by tuning from one broadcast station to another. Thus if the original adjustment employed the 69th harmonic to give zero beat with a station of 690 kc, then by doing nothing but retuning so as to receive a station operating at 1050 kc one should find substantially zero beat between this station and the 105th harmonic. It is usually found that the better class stations will agree with each other to within a very few cycles. The foregoing procedure for checking the accuracy is always desirable since it avoids the possibility of adjusting the wrong harmonic to zero beat with a broadcast station. Thus if in the preceding case one had, by mistake, adjusted the oscillator so that the 68th harmonic had a frequency of 690 kc, then the oscillator frequency would have been 10,147+ cycles, and there would be no harmonic that would zero beat with other broadcast stations.

A complete assembly for making use of broadcast stations in frequency measurement would consist of a 10-kc tuned-circuit type of oscillator of high frequency stability (see Chap. XII), a buffer amplifier, a 10-kc multivibrator, a broadcast receiver, an oscillating-detector type of receiver with plug-in coils, and a heterodyne frequency meter. This equipment can be mounted on a relay rack and represents an inexpensive frequency assembly which has an accuracy sufficient for all laboratory uses except those requiring a primary standard. The accuracy is fully as great as that of ordinary secondary standards since one normally has a number of broadcast stations continuously available for

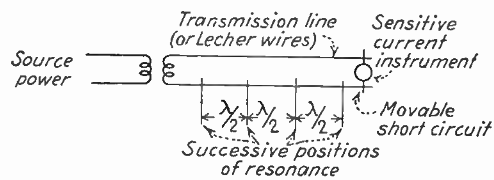



FIG. 77.—Lecher wire arrangement for directly measuring wave length at very high frequencies.

checking the accuracy of the 10-kc oscillator. By placing the tuned circuits of the 10-kc oscillator (but not the oscillator tube itself) in a box lined with balsa wood, rapid temperature changes will be avoided and very little readjustment will be needed over appreciable periods of time.

**32. Frequency Measurements with Lecher Wires.**—At very high frequencies the wave length may be determined directly by measurements made of the standing wave patterns on a resonant transmission line. The fundamental idea is illustrated in Fig. 77, which shows a simple transmission line coupled at one end to a source of power whose frequency is to be measured, and short-circuited at the other by a movable bridge in series with which is a sensitive thermocouple instrument. As the position of the short-circuiting bridge is varied, a series of sharply defined positions will be found for which the transmission line is of the proper length to give resonance, as indicated by a large current through the thermocouple instrument.[1] From the theory of

[1] In making measurements of frequency in this way it is sometimes found that the expected sharp maximum of current is replaced by a broad maximum or even a double-humped maximum. This behavior results from coupling between the parts of the transmission line on the two sides of the

transmission lines, it can be readily demonstrated that these positions of resonance are exactly one-half wave length apart. The frequency is then determined by the wave length thus obtained according to the equation

$$f = \frac{300{,}000{,}000}{\lambda} \tag{36}$$

where $\lambda$ is the wave length in meters, and $f$ is in cycles. A modification of this technique consists in replacing the short-circuiting bridge shown in Fig. 77 by a large metal disk with holes through which the wires pass, thus shielding the unused portion of the line from the part carrying current. Resonance in the transmission line is then determined by the response of a vacuum-tube voltmeter or a second Lecher wire system loosely coupled to the active part of the transmission line.[1]

Measurements of wave length made with a Lecher wire system have about the same accuracy as is obtainable with wavemeters, and may reach 0.1 per cent if great care is employed. The chief value of this method of measuring frequency is that it can be readily employed at extremely high frequencies where ordinary measuring methods can be applied only with considerable difficulty.[2]

**33. The Multivibrator.**—The multivibrator is of such great importance in the measurement of frequencies as to warrant a special discussion. The multivibrator is a two-stage resistance-

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short-circuiting bridge, and can be avoided either by associating a shield of considerable diameter with the movable bridge or by placing additional short circuits across the unused portion of the wires in order to prevent resonance in them. See Eijiro Takagishi, On a Double Hump Phenomenon of Current through a Bridge across Parallel Lines, *Proc. I.R.E.*, vol. 18, p. 513, March, 1930.

[1] Another modification which gives extremely accurate determinations of wave length is described by Hoag. See J. Barton Hoag, Measurement of the Frequency of Ultra-radio Waves, *Proc. I.R.E.*, vol. 21, p. 29, January, 1933.

[2] For further information see Francis N. Dunmore and Francis H. Engel, A Method of Measuring Very Short Radio Wave Lengths and Their Use in Frequency Standardization, *Proc. I.R.E.*, vol. 11, p. 467, October, 1923. Also see August Hund, Correction Factor for the Parallel Wire System Used in Absolute Radio-frequency Standardization, *Proc. I.R.E.*, vol. 12, p. 817, December, 1924.

coupled amplifier in which the voltage developed by the output of the second tube is applied to the input of the first tube as shown in Fig. 78. Such an arrangement will oscillate because each tube produces a phase shift of 180°, thereby causing the output of the

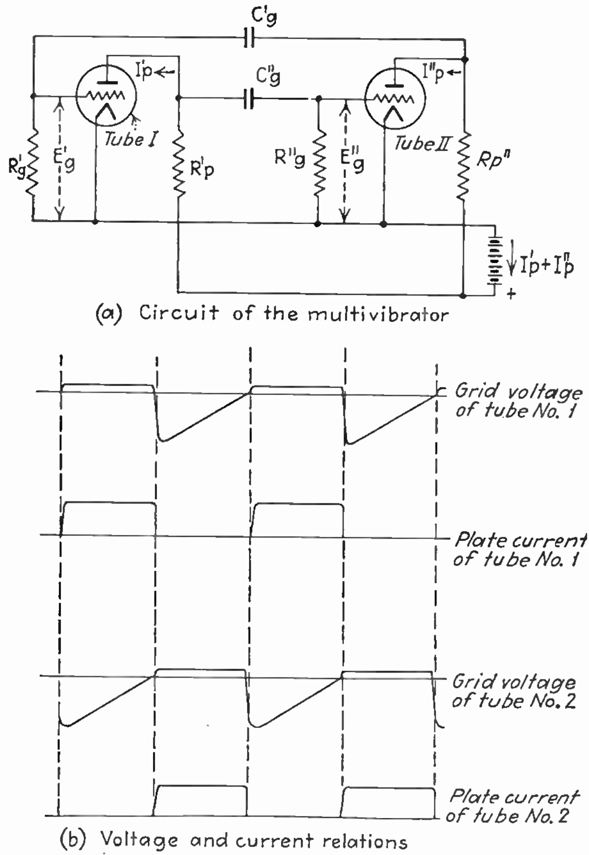



FIG. 78.—Circuit of the multivibrator, together with oscillograms showing the way in which the instantaneous grid potential and plate currents vary during the cycle of operation.

second tube to supply an input voltage to the first tube that has exactly the right phase to sustain oscillations. The usefulness of the multivibrator arises from the fact that the wave that is generated is very rich in harmonics, and from the fact that the frequency of oscillations is readily controlled by an injected voltage.

The operation of the multivibrator can be understood by reference to the oscillograms shown in Fig. 78*b*. Oscillations are started by a minute voltage at the grid of one of the tubes, say a positive potential on the grid of Tube I. This voltage is amplified by the two tubes and reappears at the grid of the first tube to be reamplified. This action takes place almost instantly and is repeated over and over, so that the grid potential of Tube I rises suddenly to a positive value, while the grid potential of Tube II just as suddenly becomes more negative than cut-off. The immediate result is then that amplification ceases, and for the moment one tube is drawing a heavy plate current while the other tube takes no plate current. This situation does not endure permanently, however, because the leakage through the grid-leak resistances gradually brings the grid potentials back toward normal. When this leakage has reached the point where amplification is just on the verge of being possible, some minute voltage will change the potentials enough to start the amplification process in the reverse direction; *i.e.*, the grid of Tube I will suddenly become negative and the grid of Tube II positive. This action is clearly evident in the oscillograms and is exactly the same as the initial action except that the relative functions of the two tubes have been interchanged. Next the potentials on the two grids gradually die away, as the result of the action of the grid leaks just as before, and finally reach a point at which the cycle repeats.

The frequency of the multivibrator oscillation is determined primarily by the grid-leak resistance and grid-condenser capacity, but is also influenced by the remaining circuit constants, the tube characteristics, and the electrode voltages. The time required to complete one cycle is proportional to the sum of the time constants of the grid-leak and grid-condenser combinations of the two tubes; *i.e.*, $1/f$ is proportional to $R_g'C_g' + R_g''C_g''$. In making preliminary estimates of the circuit constants required to give any particular frequency, one should make the sum of the time constants equal to the reciprocal of the desired frequency.

The multivibrator can be adjusted to generate frequencies ranging anywhere from perhaps 1 cycle per minute to frequencies in excess of 100,000 cycles per second. The upper limit is the highest frequency at which satisfactory resistance-coupled amplification is possible, while the lower limit is fixed by the leakage of the grid condenser in relation to the condenser capacity.

It will be observed that the multivibrator oscillations are far from sinusoidal, and in particular have sharp corners that indicate high-order harmonics. It is found in practice that detectable harmonics beyond the three hundredth are commonly present in the output of a multivibrator. This ability to produce harmonics is one of the reasons the multivibrator is found to be so useful in frequency-measuring work.

*Synchronization with Injected Voltage.*—Injection into the multivibrator circuit of a voltage having a frequency higher than the multivibrator oscillation tends to cause the latter to adjust itself to a frequency which is $1/n$ of the injected frequency, where $n$ is an integer. It is possible in this way to use the multivibrator to reduce frequency (*i.e.*, to generate a subharmonic of the injected frequency), and it is entirely practicable to maintain the control rigidly when $n$ is as large as 10.

The exact details of the control depend upon the way in which the synchronizing voltage is injected, and upon the amplitude of the synchronizing potential. Typical cases are shown in Fig. 79. It will be observed in all cases that the injected voltage causes the polarity of either one or both of the grids to reverse sooner than would otherwise be possible. This comes about because the superimposed voltage can momentarily neutralize charges that are trapped on grid condensers and thereby cause the polarity to reverse without waiting for these charges to leak away.

Increasing the amplitude of the synchronizing frequency causes the multivibrator frequency to be "drawn" in discontinuous steps toward the synchronizing frequency, with the ratio being expressible always as an integer of progressively smaller magnitude. This action is brought out very clearly by the experimental results presented in Fig. 80, and is caused by the fact that increasing the amplitude of the injected voltage enables it to neutralize larger condenser charges, and hence to cause a reversal of grid potential before the full charge has had time to leak away.

The way in which the multivibrator frequency is drawn toward the synchronizing frequency depends upon where this

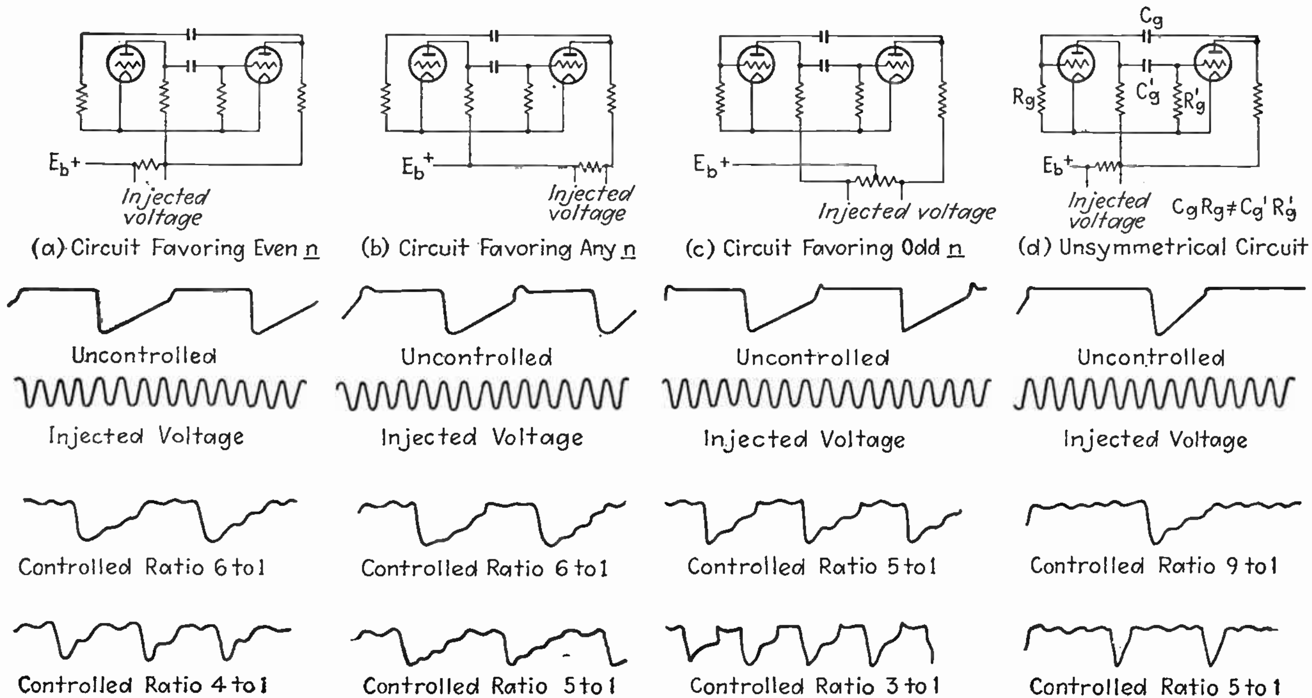



FIG. 79.—Oscillograms showing the mechanism whereby an injected voltage is able to control the frequency of the multivibrator oscillations.

voltage is injected, and upon the degree of symmetry between the circuit constants of the two multivibrator tubes. With perfect symmetry the arrangement shown at Fig. 79*a* favors even values of *n* because it treats both tubes alike. Thus if the synchronizing voltage is increased until the positive grid period of one tube is shortened by one cycle of the synchronizing frequency, the positive grid period of the other tube will likewise be shortened by the same amount, and the multivibrator cycle will be shortened by two cycles of the synchronizing frequency. In contrast to this, injection of the synchronizing potential into only one plate circuit tends to cause the shortening action to be confined to one tube, so that the multivibrator cycle is shortened in steps corresponding to one cycle of the synchronizing frequency. Both even and odd values of *n* are then obtainable as shown in Fig. 79*b*. Again, if the synchronizing voltage is introduced in phase opposition in the two plate circuits of a symmetrical multivibrator as shown at Fig. 79*c*, the result favors odd values of *n*. Finally, if the multivibrator circuit is unsymmetrical, the successive periods of the multivibrator cycle have unequal lengths, and depending upon the ratio of these lengths either even or odd values of *n* can be favored irrespective of whether the circuit of Fig. 79*a* or Fig. 79*c* is used. This is shown at Fig. 79*d*.

The effect of the synchronizing voltage amplitude upon the frequency of the multivibrator oscillations is shown in Fig. 80 for various methods of injection. The very strong tendency to favor even values of *n* in some cases, and odd values in other instances, is clearly evident even though small dissymmetries were present.

*Construction and Operation of Multivibrators.*—The design and construction of a multivibrator circuit should follow the same principles as those employed in resistance-coupled amplifiers, except that the grid condenser should have a capacity such that the desired frequency will be generated when the grid-leak resistance is of the value that would normally be used in resistance-coupled amplification. Where high-order harmonics are sought, care must be taken in the design and layout to insure that the amplifier has the best possible high-frequency response consistent with reasonable amplification. This means that the capacities in shunt with the coupling resistance must be as low as possible, and makes screen-grid and radio-frequency pentode tubes particularly suitable because of their low input capacity.

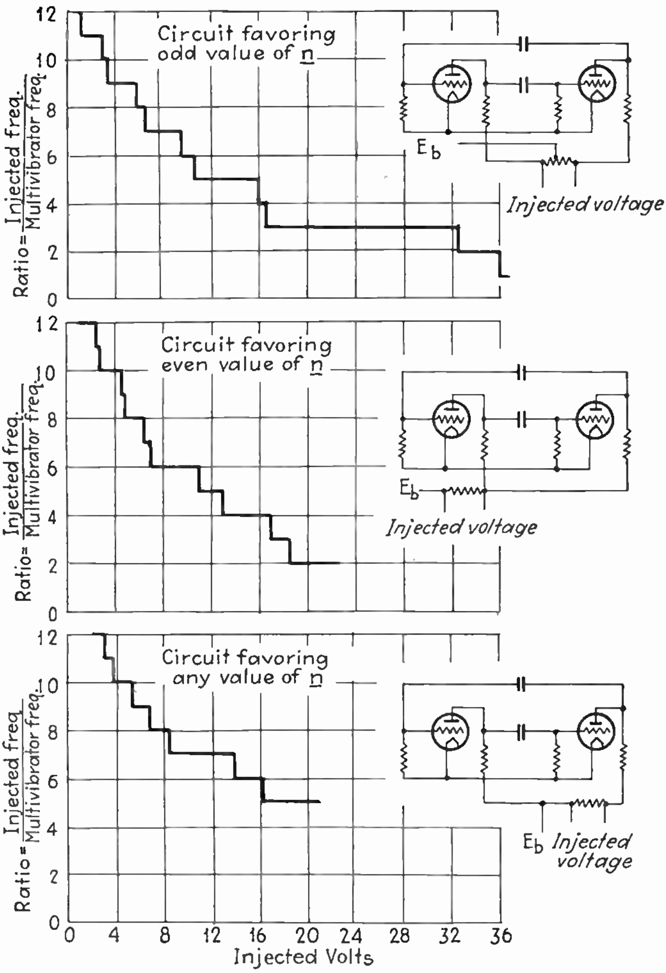



Fig. 80.—Experimental results showing the effect of increasing the amplitude of the synchronizing voltage, and how certain methods of injecting the voltage tend to favor either even or odd frequency ratios.

When the multivibrator is to be controlled, as is normally the case, the uncontrolled frequency of oscillation should be slightly less than the value when controlled, and in order to obtain the maximum stability of the control the injected frequency must have the proper amplitude, and the circuit arrangements should be such as to favor strongly either even or odd values of $n$ as desired.

**34. Measurement of Audio Frequencies.**—Audio frequencies are commonly measured by comparing the unknown frequency with a known frequency, or by balancing a bridge of a type in which the conditions of balance depend upon the frequency.

Any bridge in which the balance depends upon frequency can be used to measure frequency in terms of the circuit constants of the bridge. The Hay,[1] resonance, and Wien bridges described in Chap. II are of this type. The Wien bridge is the most satisfactory for frequency measurements, however, since it does not require an inductance, can be brought into balance merely by varying resistance elements, and is readily proportioned to cover a very wide frequency range. The absence of an inductance in the bridge arms is important since the large inductances required at audio frequencies may pick up energy from stray fields and introduce errors in the balance.

A Wien bridge arranged for measuring frequency is shown in Fig. 81. Balance in such a bridge is obtained by simultaneously satisfying the two equations

$$f = \frac{1}{2\pi \sqrt{R_c R_d C_c C_d}} \tag{37a}$$

and

$$\frac{C_d}{C_c} = \frac{R_b}{R_a} - \frac{R_c}{R_d} \tag{37b}$$

By making $C_c = C_d$, $R_c = R_d$, and $R_b/R_a = 2$, the second condition of balance is always satisfied, and one has

$$f = \frac{1}{2\pi R_c C_c} \tag{38}$$

Thus by making $R_c$ and $R_d$ identical slide-wire resistances, and mounting them on a common shaft, the dial can be calibrated directly in frequency. Furthermore, convenient multiplying factors, such as decimal values, can be obtained by merely

[1] The use of the Hay bridge for frequency measurements is discussed in detail by Chester I. Soucy and B. de F. Bayly, A Direct Reading Frequency Bridge for the Audio Range Based on Hay's Bridge Circuit, *Proc. I.R.E.*, vol. 17, p. 834, May, 1929.

changing the capacities of the condensers $C_c$ and $C_d$ by the appropriate amounts. A frequency range of 10 to 1 can be readily covered with a single condenser value, so that by the use of three pairs of condensers one can cover the complete audio range from 20 to 20,000 cycles. In practical construction it is impossible to maintain equality between $R_c$ and $R_d$ with the accuracy required to maintain a perfect balance, so that the potentiometer $P$ having a total resistance of perhaps 1 to 2 per cent of $R_a$ is used to sharpen the balance. This has no effect upon the frequency calibration.

An accuracy in the order of ½ to 1 per cent can be readily obtained in frequency bridges. In the frequency range 300 to 5000 cycles, balance can be most easily made by the use of telephone receivers. At frequencies outside this range some indicating arrangement, such as an amplifier and vacuum-tube voltmeter combination, is normally required. The principal difficulty involved in making accurate frequency measurements by bridge methods arises from harmonics of the frequency being measured. The bridge is unbalanced for these harmonics, which are thus very prominent in the bridge output even though only a small percentage in the bridge input current. When telephone receivers are used in the middle audio range, it is usually possible to balance the bridge for the fundamental in spite of the presence of harmonics; but when indicating instruments are employed, appropriate filters must be placed in the output of the neutral arm to prevent spurious voltages from reaching the indicating device.

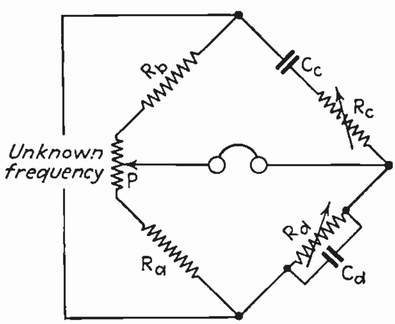


FIG. 81.—Wien bridge arranged for the measurement of frequency.

*Comparison Methods of Measuring Audio Frequencies.*—Standard frequencies for use in audio-frequency measurements may be obtained in a number of ways. Thus the synchronous clock of a primary standard is normally operated from 1000-cycle power, and this is a convenient audio-frequency standard. Electrically driven tuning forks represent another source of standard audio frequencies. Where known radio frequencies are available, it is always possible to adjust a low-frequency oscillator in a known relation to the high-frequency standard by the use of harmonics. In the case of the low audio frequencies it is also possible to measure the frequency directly by means of synchronous clocks or other methods. In this connection it is to be noted that many ordinary clocks designed for use at 60 cycles will operate synchronously over a frequency range extending from about 30 cycles to over 100 cycles.

With one or more known audio frequencies available it is possible to calibrate completely an audio-frequency oscillator by the use of harmonics. Thus if the known frequency is 100 cycles, one can obtain a calibration point for every harmonic of 100 cycles either by comparing known and unknown frequencies directly with a cathode-ray oscillograph, as described in Sec. 79, or by zero beating harmonics of the standard with the unknown. If the standard frequency is a high value, such as 1000 cycles, one can set up an auxiliary oscillator to exactly 100 cycles by zero beating its tenth harmonic with the standard. Many modifications of these harmonic methods are possible and can be readily worked out to fit any particular set of circumstances.

Two audio frequencies can be adjusted to zero beat with each other, *i.e.*, to identical frequencies, by combining the outputs in a telephone receiver to give aural beats, by applying the two frequencies simultaneously to a vacuum-tube voltmeter and observing the vibrations of the meter pointer, or by use of a cathode-ray oscillograph as discussed in Sec. 79. The particular method preferred in any individual case depends upon the circumstances involved.

An audio oscillator after being calibrated against a standard frequency can then be used as a source of known frequencies for calibrating other equipment. For greatest accuracy the oscillator should be of the tuned-circuit type, but the beat-frequency type of oscillator, although not holding its calibration so well, is easier to use because it will sweep continuously through its entire frequency range with the turn of a single dial.

The accurate comparison of an unknown frequency with known fixed frequencies involves special problems. The usual procedure is to obtain beats representing the difference between the unknown frequency and the nearest available known frequency. The difference frequency so obtained can then be determined by counting the beats by visual or aural methods if the frequency does not exceed a few cycles per second, and by either mechanical or electrical counting for higher frequency beats.[1]

Audio frequencies are somewhat more difficult to measure with high accuracy than are radio frequencies. If a frequency of 100 cycles is to be measured with a precision of one part in ten million, this is equivalent to one cycle error every 28 hr. It thus takes a long time to make an accurate determination unless a high harmonic of the audio frequency is compared with a radio-frequency standard. The accuracy normally required in audio-frequency measurements is fortunately less than the precision commonly necessary at radio frequencies, being usually in the order of 1 to 0.1 per cent.

**35. Determination of Small Percentage Changes in Frequency.** There is an important group of measurements which involve the accurate determination of small frequency differences. Typical examples are supplied by the frequency-variation method of measuring radio-frequency resistance, where frequency changes of the order of 1 per cent must be determined, and by the method of measuring small capacities by the change they produce in the frequency of an oscillator, where frequency changes of only a few cycles must sometimes be determined.

The customary method of making measurements of this type is illustrated schematically in Fig. 82. An auxiliary oscillator of good frequency stability over short time intervals is adjusted until its frequency differs by a convenient amount from the frequency of the oscillator under test, and some means, commonly comparison with an audio-frequency oscillator, is used to measure the difference frequency produced when the two radio-frequency oscillations are applied to a detector. If the frequency of the oscillator under test now changes by a small amount, even only two or three cycles, the audio-frequency beat note will change by the same number of cycles, and the difference can be

[1] Examples of such methods are given by N. P. Case, A Precise and Rapid Method of Measuring Frequencies from 5 to 200 Cycles per Second, *Proc. I.R.E.*, vol. 18, p. 1586, September, 1930; F. Guarnaschelli and F. Vecchiacchi, Direct Reading Frequency Meter, *Proc. I.R.E.*, vol. 19, p. 659, April, 1931.

accurately determined because it represents a comparatively large percentage change of the audio frequency.

In making the measurements, the auxiliary oscillator is preferably adjusted to give an audio-frequency beat note with the oscillator under test, rather than adjusted to zero beat. This avoids all possibility of errors from automatic synchronization between the two oscillators and makes it possible to determine whether the frequency has increased or decreased. Care must be

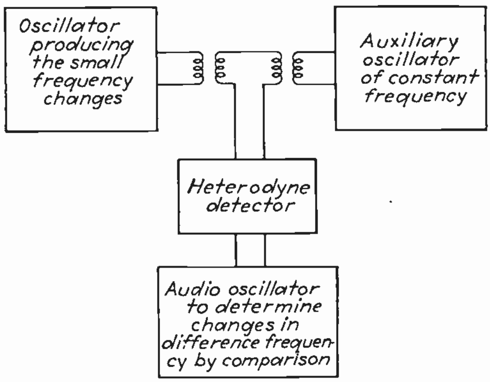



FIG. 82.—Schematic diagram illustrating how small change in frequency may be accurately measured.

taken, however, to arrange matters so that the zero-beat condition is not passed through as the frequency changes. The accuracy of the measurements rests upon the assumption that the frequency of the auxiliary oscillator does not change during the measuring operation. If the time involved is only a minute or so, ordinary tuned-circuit oscillators properly designed will normally be satisfactory. It is usually possible to combine the auxiliary oscillator and the detector into one unit in the form of an oscillating detector. When the accuracy requirements are very severe, or when long periods of time are involved, it may be necessary to use a crystal-controlled oscillator.

## CHAPTER VI

# WAVE FORM AND PHASE

**36. Wave Shapes of Audio-frequency Waves by Oscillographic Methods.**—Wave shapes at the lower audio frequencies can be readily observed visually or photographed for permanent record with the aid of an ordinary magnetic oscillograph. The frequency range that can be covered in this way is limited, however, by the resonant frequency of the vibrating element and is usually of the order of 1500 to 3500 cycles, with appreciable phase distortion beginning at frequencies perhaps one-third of these.

It requires about 100 ma to give a deflection of 1 in. on the usual magnetic oscillograph, and this often makes amplification necessary. A high-grade push-pull amplifier with an output transformer having the proper turn ratio to match the resistance of the oscillograph element to the power tubes is fairly satisfactory but introduces some phase shift and some frequency discrimination at low and high frequencies because of the imperfections of the transformer. Where the oscillograph amplifier is to be used over a wide frequency range and where the distortion introduced must be negligible, a special resistance-coupled push-pull amplifier with a direct-coupled output as shown in Fig. 83 is required. This arrangement provides push-pull power amplification with no transformers through the expedient of a phase-reversing tube to excite one side of the push-pull circuit. It is possible in this way to obtain almost any desired frequency range with negligible amplitude, frequency, or phase distortion, provided proper circuit constants are used. A high plate-supply voltage $E_b$ is desirable since this will allow resistances $R_1$ and $R_2$ to be large, thus tending to straighten out the dynamic characteristic of the power tubes and so reduce distortion. It is generally necessary to use two or three tubes on each side of the push-pull output stage to obtain sufficient output capacity. When this is the case, resistances should be placed in series with the grid of each individual output tube to prevent parasitic oscillations which can otherwise be expected.

A cathode-ray oscillograph with a linear time axis can be used instead of a magnetic oscillograph if desired. The cathode-ray tube has the advantage of practically no frequency limitation, and also consumes negligible power if electrostatic deflection is employed. The production of a time axis involves complications, however, and difficulties are also encountered in timing the camera shutter when photographing transients.

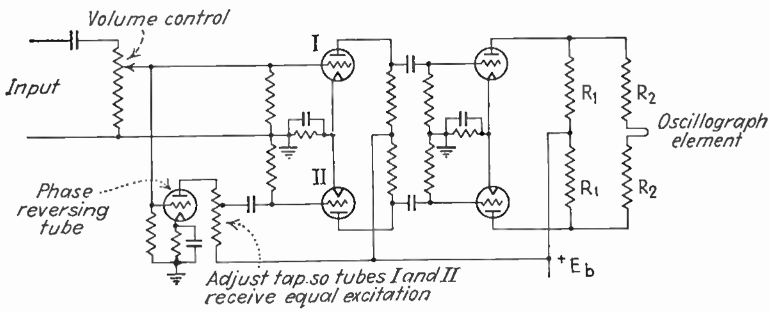



FIG. 83.—Push-pull oscillograph amplifier having no transformers.

**37. Harmonic Analysis of Oscillograms.**—Periodic electrical waves can always be expressed as the sum of a direct-current component plus a series of harmonics which are all harmonics of the fundamental frequency of the wave. That is, if $E$ is a periodic function, one can then write

$$E = B_0 + C_1 \sin(\omega t + \Phi_1) + C_2 \sin(2\omega t + \Phi_2) + C_3 \sin(3\omega t + \Phi_3) + \cdots \quad (39a)$$

where

$B_0$ = amplitude of direct-current component
$C_1$, $C_2$, etc. = amplitudes of corresponding alternating-current components
$\Phi_1$, $\Phi_2$, etc. = phase angle of corresponding alternating-current components
$\omega$ = $2\pi$ times the fundamental frequency of the wave.

Equation (39a) is often rearranged as

$$E = A_1 \sin \omega t + A_2 \sin 2\omega t + A_3 \sin 3\omega t + \cdots + B_0 + B_1 \cos \omega t + B_2 \cos 2\omega t + B_3 \cos 3\omega t + \cdots \quad (39b)$$

where

$$\sqrt{A_n^2 + B_n^2} = C_n$$
$$\frac{B_n}{A_n} = \tan \Phi_n$$

*Fourier Method of Determining the Coefficients of an Arbitrary Curve.*—The $A$ coefficients appearing in Eq. (39*b*) can be evaluated by making use of the fact that, when a periodic curve is multiplied by sin $n\omega t$ and the resulting area summed up over one complete cycle of the fundamental frequency, the result is determined only by the coefficient $A_n$ in Eq. (39*b*). Likewise, when the curve is multiplied by cos $n\omega t$ and the resulting area summed up over a complete cycle, the result is determined solely by the coefficient $B_n$. By making use of this principle one obtains the following formulas:[1]

$$A_n = \frac{1}{\pi}\int_{\omega t = 0}^{\omega t = 2\pi} E \sin n\omega t \, d(\omega t) \tag{40a}$$

$$B_n = \frac{1}{\pi}\int_{\omega t = 0}^{\omega t = 2\pi} E \cos n\omega t \, d(\omega t) \tag{40b}$$

$$B_0 = \frac{1}{2\pi}\int_{\omega t = 0}^{\omega t = 2\pi} E \, d(\omega t) \tag{40c}$$

where $E$ represents the curve being analyzed.

The integration indicated in Eqs. (40) can be carried out mathematically when the wave being analyzed follows some known law, as, for example, when the wave analyzed is the voltage output wave of a rectifier. In other cases the product under the integral sign can be calculated point by point, plotted, and the resulting net area (positive minus negative areas) determined by counting squares, by use of a planimeter, or with the aid of Simpson's rule. Ordinary harmonic analyzer machines make use of Eqs. (40), and are simply devices for multiplying the curve being analyzed by the proper sine or cosine function, and then summing up the resulting area over a complete cycle. The Fourier method gives directly the correct amplitude of any particular component irrespective of the presence or absence of other components.

[1] The development of these equations is to be found in every mathematical discussion of Fourier series, but is beyond the scope of this book.

*Schedule Method.*—In the schedule method one cycle of the arbitrary curve which is to be analyzed is divided up into a number of equally spaced ordinates as shown in Fig. 84. As many coefficients as there are ordinates are then evaluated so that the resulting curve will pass through the selected points. Thus in Fig. 84 it is possible to obtain an equation that will pass through the selected ordinates by assigning proper values to $A_1$, $A_2$, $B_0$, $B_1$, $B_2$, and $B_3$ while making all other coefficients in Eq. (39*b*) equal to zero. Evaluating the coefficients to accomplish this result involves solving as many simultaneous equations as there are coefficients to be evaluated. By taking advantage of the fact that the ordinates are evenly spaced, the solution of these simultaneous equations can be simplified to the point where it

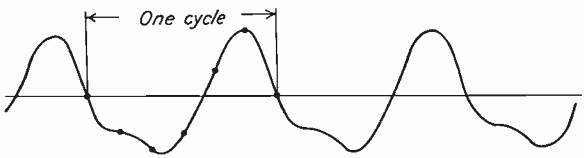



FIG. 84.—Arbitrary curve with six equally spaced ordinates. A six-point schedule will give a curve that will pass through the six selected points, and this calculated curve will or will not coincide with the actual curve between selected points according to whether or not the actual curve contains coefficients not evaluated by the six-point schedule.

involves performing only a few simple multiplications, and carrying out a series of simple additions and subtractions. These operations are commonly indicated by a form or schedule—hence the name "schedule method."[1] A simple schedule for the case of six ordinates is given in the accompanying table. Schedules for 12-, 18-, and 36-point analyses are available for both the general case where there is a direct-current component and both even and odd harmonics, and also where only odd harmonics are present. These schedules with more ordinates are naturally more complicated than the one shown in the accompanying table, but will likewise evaluate a greater number of coefficients.

The schedule method evaluates the coefficients so that the resulting curve is correct for the selected ordinates. Between

[1] The method of deriving schedules is described by F. W. Grover, Analysis of Alternating-current Waves by the Method of Fourier, with Special Reference to Methods of Facilitating Computations, *Reprint* 203 from *Bull. of Bur. of Std.*, vol. 9, 1913. This publication gives a number of schedules.

these points the computed and actual curves will not agree, however, unless the coefficients evaluated are the only coefficients actually present in the wave. Thus in Fig. 84 the schedule method will not give the correct result if there is some fourth harmonic present. It is therefore necessary to use enough

Table VI.—Six-point Schedule Involving Both Even and Odd Harmonics and a Constant Term

| Measured ordinates | | Sums | Differences | | Sine terms: $A_1$ and $A_2$ | | Cosine and constant terms: $B_1$ and $B_2$ | | $B_0$ and $B_6$ | |
|---|---|---|---|---|---|---|---|---|---|---|
| $y_0$ | | $s_0$ | $d_0$ | | | | | | | |
| $y_1$ | $y_5$ | $s_1$ | $d_1$ | sin 30° | ............. | | $-s_2$ | $s_1$ | | |
| $y_2$ | $y_4$ | $s_2$ | $d_2$ | sin 60° | $d_1$ | $d_2$ | | | | |
| $y_3$ | | $s_3$ | $d_3$ | sin 90° | ............. | | $s_0$ | $-s_3$ | $s_0 + s_2$ | $s_1 + s_6$ |
| | | | | Sums | $S_0'$ | $S_e'$ | $S_0''$ | $S_e''$ | $S_0'''$ | $S_e'''$ |
| | | | | | $A_1 = \frac{S_0' + S_e'}{3}$ | | $B_1 = \frac{S_0'' + S_e''}{3}$ | | $B_0 = \frac{S_0''' + S_e'''}{6}$ | |
| | | | | | $A_2 = \frac{S_0' - S_e'}{3}$ | | $B_2 = \frac{S_0'' - S_e''}{3}$ | | $B_6 = \frac{S_0''' - S_e'''}{6}$ | |

Checks

$$s_0 = (B_0 + B_3) + (B_1 + B_2)$$
$$s_2 = 2(B_0 + B_3) - (B_1 + B_2)$$
$$s_0 + s_2 = 3(B_0 + B_3) \qquad 2s_0 - s_2 = 3(B_1 + B_2)$$
$$s_1 = 2(B_0 - B_3) + (B_1 - B_2)$$
$$s_3 = (B_0 - B_3) - (B_1 - B_2)$$
$$s_1 + s_3 = 3(B_0 + B_3) \qquad s_1 - 2s_3 = 3(B_1 - B_2)$$
$$d_1 = 2(A_1 + A_2) \sin 60°$$
$$d_2 = 2(A_1 - A_2) \sin 60°$$

*Procedure.*—The measured ordinates are first written down in two columns in the order indicated. In the next two columns appear the sums $s_m$ of the ordinates, found by adding those in the same row, and the differences $d_m$ of the same ordinates. In the fifth column are indicated the trigonometric functions which enter into the calculation. The rest of the schedule indicates in an abbreviated form what products are to be formed, the convention being adopted that each quantity $s_m$ or $d_m$ is to be multiplied by the sine of the angle which appears in the same row at the left. Thus one forms the product $s_1 \sin 60°$ in one case, of $-s_2 \sin 30°$ in another case, and $(s_1 + s_3) \sin 90°$ in still another.

ordinates to permit the evaluation of all important frequency components present in the wave.

**38. Harmonic Analyzers.**—A variety of experimental methods have been devised for analyzing complex voltage and current waves. Some of the most practical of these are described below.

*Dynamometer Method.*[1]—In this method the wave to be analyzed is passed through one coil of a dynamometer instrument while a search current of controllable frequency is passed through the other coil. The operation of the device depends upon the fact that the instrument pointer will not be deflected unless the frequency of the search current is equal to or very close to the frequency of a component of the wave being analyzed. When the difference between the two frequencies is a fraction of a cycle per second, the pointer will pulsate at the difference frequency with an amplitude of pulsation equal to $I_n I_s$, where $I_n$ is the effective amplitude of the unknown component (corresponding to coefficient $C_n$ in Eq. 39), and $I_s$ is the effective amplitude of the search current. Thus to analyze a wave, one varies the frequency of the search current and notes the frequencies at which beats occur. Each appearance of beats indicates a frequency component in the unknown wave equal to the search frequency, and having an amplitude proportional to the amplitude of the beats.

This type of frequency analyzer can be calibrated on direct currents, since the deflection of the instrument is the same when $I_n$ and $I_s$ are direct currents as when they are effective values of alternating currents. Another method of calibration which is particularly useful at high frequencies is to substitute for the complex wave a known voltage of the desired frequency, and then to adjust this voltage until the amplitude of the beats is the same as obtained with the unknown wave. If the search current $I_s$ is maintained constant throughout, accurate results can be obtained even when the normal frequency range of the dynamometer instrument is greatly exceeded.

The dynamometer method of analyzing a complex wave has the merits of simplicity and directness. The ability of the method to measure accurately small harmonic components in the presence of large components of other frequencies is determined by the allowable power dissipation in the dynamometer coils, since this limits the current that can be passed through the coils and hence the amplitude of the beats. With practical instruments of the proper sensitivity there is no difficulty in

[1] M. G. Nicholson and William M. Perkins, A Simple Harmonic Analyzer, *Proc. I.R.E.*, vol. 20, p. 734, April, 1932.

measuring the distortion commonly encountered in power amplifiers used in radio receivers.

*Vacuum-tube Voltmeter Method of Harmonic Analysis.*[1]—This method is illustrated in Fig. 85 and involves superimposing a search voltage upon the wave being analyzed and then applying the resulting wave to a *full-wave square-law* vacuum-tube voltmeter. In such an arrangement the vacuum-tube voltmeter will give a steady deflection depending only upon the effective value of the combined wave. When the search voltage has a frequency within a fraction of a cycle of some frequency component contained in the unknown wave, then beats are superimposed upon

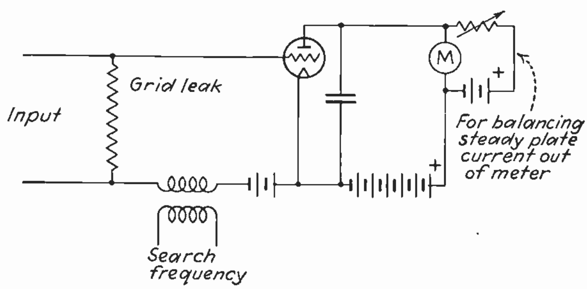



FIG. 85.—Vacuum-tube voltmeter method of analyzing wave forms. If desired, a shielded input transformer may be used to couple the wave being measured to the vacuum-tube voltmeter.

this steady deflection. These beats have a crest amplitude proportional to $E_nE_s$, where $E_s$ is the crest amplitude of the search frequency voltage, and $E_n$ is the crest amplitude of the frequency component of the unknown wave that differs from the search frequency by only a fraction of a cycle. One can thus measure the amplitude and frequency of each component of the unknown wave by varying the search frequency, and noting the frequencies at which beats occur and the amplitude of the beats.

This type of harmonic analyzer is most satisfactorily calibrated by the substitution method, *i.e.*, by substituting for the unknown wave an adjustable known voltage which gives the same amplitude of beats as developed by the unknown wave, and has the same frequency.

A full-wave square-law vacuum-tube voltmeter is absolutely essential if accuracy is to be obtained, and the input voltages

[1] See Chauncey Guy Suits, A Thermionic Voltmeter Method for the Harmonic Analysis of Electrical Waves, *Proc. I.R.E.*, vol. 18, p. 178, January, 1930.

applied must be restricted to the range over which the tube voltmeter possesses this characteristic. Otherwise the amplitude of beats is not necessarily proportional to the amplitude of the component being measured.

In order to obtain maximum sensitivity, the steady deflection upon which the beats are superimposed should be partially or completely balanced out of the plate meter by one of the usual methods. It is to be noted that, when the device is being calibrated by the substitution method, the steady deflection will change when the locally generated known voltage is substituted for the complex wave. This introduces no uncertainty, since with a full-wave square-law detector the amplitude of the beats is the significant thing, and the steady deflection is of no consequence.

The vacuum-tube voltmeter type of harmonic analyzer has the same merits of simplicity and directness as the dynamometer method, and has the further advantage of consuming negligible power. The ability to measure small harmonic components in the presence of large components of other frequencies is limited, first, by the fact that the total crest voltage applied to the vacuum-tube voltmeter (sum of unknown and search voltages) must be limited to the square-law part of the tube characteristic; and, second, by the sensitivity of the plate meter of the vacuum-tube voltmeter. With a 500-$\mu$a. meter there is no difficulty in measuring accurately the distortion in the output of a power amplifier tube.

*Resonance-bridge Method of Harmonic Analysis.*—The total r.m.s. value of harmonics contained in a current wave can be measured by means of the bridge circuit shown in Fig. 86, which consists of two resistance arms and two series resonance arms. These resonant arms must be so proportioned that at the fundamental frequency $\omega L/R > 10$. The resistance of the potentiometer $P$ across the neutral arm is normally made equal to the resistance $R$.[1]

The procedure for measuring the harmonic content of a wave is as follows: The bridge is first balanced at the fundamental frequency by varying the $LC$ product and the series resistance

[1] A slightly different form of resonance bridge is described by D. F. Schmit and J. M. Stinchfield, Measuring Harmonic Distortion in Tube Circuits, *Electronics*, vol. 1, p. 79, May, 1930.

of the resonant arms. This can be done by closing switches $S_1$ and $S_2$ and temporarily connecting a telephone receiver across the voltage divider $P$. The voltage or current to be analyzed is then applied to the bridge, and with switches $S_1$ and $S_2$ open the voltage divider $P$ is adjusted to give a convenient reading on the meter $M$. The voltage across $P$ under these conditions is the r.m.s. potential developed by the unknown current wave in flowing through the resistance $P$. Switches $S_1$ and $S_2$ are now closed, and the setting of $P$ varied until the same indication is obtained on meter $M$ as before. The voltage across $P$ is now the potential developed by the r.m.s. value of the harmonics of the

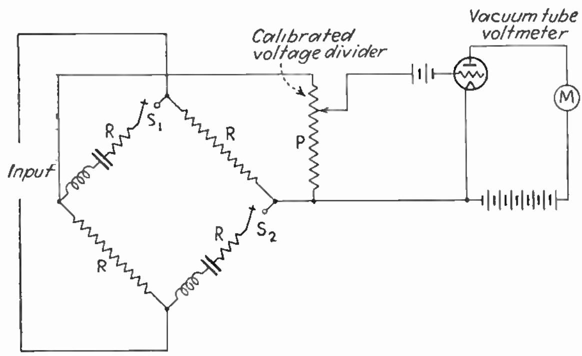



FIG. 86.—Resonance bridge for measuring r.m.s. harmonic content of a wave.

current wave as they flow through the resistance of $P$. This is because the bridge is balanced to the fundamental, so no fundamental voltage appears across $P$, while the resonant arms offer such high impedance at other than the fundamental frequency that practically all the harmonic current is forced to flow through the neutral arm. The ratio of r.m.s. harmonic content to total r.m.s. value of the wave is then the ratio of the two settings of the voltage divider $P$.

This method of harmonic analysis is capable of measuring very small harmonics in the presence of a very large fundamental, and can be used to determine distortion in audio-frequency power amplifiers. It is not capable of determining the amplitudes of the individual frequency components, however, and becomes cumbersome when measurements are to be made at more than one fundamental frequency. The impedance which the bridge offers also varies, being $2R + P$ to all frequencies when switches

$S_1$ and $S_2$ are open, and being $R$ to the fundamental and $2R + P$ to the harmonics when the switches are closed.

*Heterodyne Method of Harmonic Analysis.*[1]—In the heterodyne method of wave analysis the component to be measured has its frequency increased to a predetermined value by heterodyne action, and is then amplified and measured at this fixed frequency. The amplitude of the unknown component is obtained from the measured output by means of a suitable calibration, while its frequency is determined from the heterodyne frequency that must be used to change its frequency to the known fixed frequency.

A typical circuit for carrying out the necessary operations is shown in Fig. 87 and includes as essential features the crystal filter in the fixed frequency amplifier, and a balanced modulator. The balanced modulator is necessary in order to prevent heterodyne-oscillator energy from forcing its way through the fixed amplifier when the frequency component being measured is so low that the heterodyne oscillator operates within a few cycles of the fixed frequency. The balance must be made so perfect with the aid of balancing condenser $C$, potentiometer $P$, and the shielded transformers that, when the heterodyne oscillator is operated at the frequency of the fixed amplifier, the output of the latter is negligible. It is also essential that the heterodyne oscillator be thoroughly shielded from the fixed frequency amplifier. To avoid errors from high-order modulation products, the balanced modulator should be so adjusted that the instantaneous grid potential is always on the square-law part of the tube characteristic, and the input voltage should be kept as low as is consistent with satisfactory sensitivity. The crystal filter is necessary because of the extremely high selectivity required in analyzing low-frequency waves. Thus if one wishes to measure a small second harmonic of a 30-cycle fundamental, the same heterodyne frequency that changes the component to be measured to a fixed frequency 50,000 cycles will change the fundamental to 49,970 cycles. In spite of the small percentage difference, the latter component must not reach the output of the fixed amplifier

[1] Further discussion of this method is given by L. B. Arguimbau, Wave Analysis, *General Radio Experimenter*, vol. 7, p. 12, June, 1933; C. R. Moore and A. S. Curtis, An Analyzer for the Voice Frequency Range, *Bell System Tech. Jour.*, vol. 6, p. 217, April, 1927. The latter reference discusses especially the problem of avoiding false indications produced by high-order modulation products.

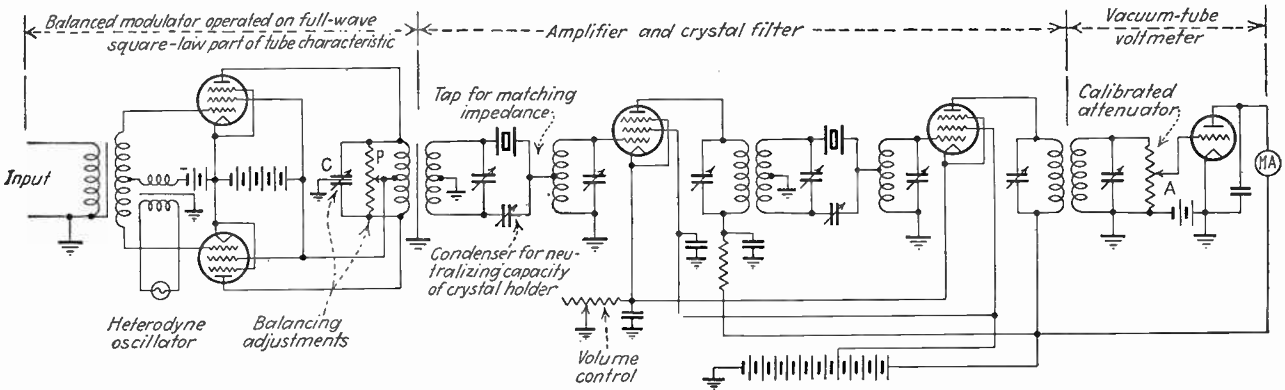



FIG. 87.—Circuit diagram of typical heterodyne type of wave analyzer.

even though it is perhaps one hundred times as large as the desired component. Two carefully matched crystals will give the required selectivity when augmented by several ordinary tuned circuits to prevent trouble from other modes of vibration.

When the equipment is arranged so that the fixed amplifier utilizes the sum of the heterodyne and unknown frequency, the frequency at which the amplifier operates must exceed twice the highest frequency to be measured in order to avoid trouble from harmonics of the heterodyne oscillator. For audio-frequency measurements this results in a fixed frequency of 40 to 50 kc, which is just right for a crystal filter. If the difference frequency is employed, the fixed frequency may be about 15 to 20 kc. Since it is difficult to obtain crystals for such a frequency, a magnetostriction type of filter may be employed.[1]

Calibration of a heterodyne type of wave analyzer is most satisfactorily carried out by the substitution method. One may use the same procedure as employed with the vacuum-tube voltmeter type of wave analyzer, but it is somewhat more satisfactory to carry out the calibrating in the following manner: A standard voltage, say 1 volt, of the frequency to be measured is applied to the input with the calibration attenuator $A$ set at a predetermined fixed value, and the volume control is adjusted until a standard output is obtained. The unknown voltage is then applied to the input and the attenuator $A$ readjusted until the standard output is again obtained. The ratio of the component being measured to the standard calibrating voltage is then given in terms of the attenuator settings, which can be made direct reading by suitable calibration. A calibration made at one frequency can be made to apply to all input frequencies provided a curve giving the necessary correction as a function of frequency is prepared.

The heterodyne wave analyzer is able to measure accurately frequency components over a very wide range of amplitude, since a large amount of amplification can be employed to take care of very small voltages, while a voltage divider associated with the input will handle large voltages. The disadvantage of the method is that the equipment is quite elaborate, and that the

[1] For a discussion of such filters see Harry H. Hall, A Magnetostriction Filter, *Proc. I.R.E.*, vol. 21, p. 1328, September, 1933.

frequency of low-frequency components is not determined accurately.[1]

*Tuned-amplifier Type of Harmonic Analyzer.*—A complex wave may be analyzed by applying it to a tuned amplifier and adjusting the resonant circuits to separate out the component to be measured so that it can be separately evaluated. This method of harmonic analysis has been developed to a high state of perfection for use in inductive interference investigations,[2] and is capable of measuring very small frequency components even when other components of large amplitude are present. The equipment required is rather elaborate, however, if a considerable frequency range is to be covered by the tuned circuits, and the manipulation is slow, particularly when the exact frequencies contained in the wave being analyzed are not known.

*Determination of Phase Position.*—The experimental methods of wave analysis described above give only the amplitude of the various frequency components and not the phase. The phase position of any frequency component of the unknown wave can be obtained by superimposing upon it a wave having the same frequency as the component being measured, and of adjustable known phase. When the phase of the injected voltage is such that a maximum resultant is obtained, the unknown and injected components are of the same phase. This method can be readily applied where the complex wave consists of a fundamental and a series of harmonics of this fundamental. The procedure is to derive from the complex wave a sine wave of the fundamental frequency by the use of a tuned amplifier, and then to apply this to a harmonic generator that is adjusted so that the angle of current flow does not exceed one-half cycle of the highest harmonic that is to have its phase determined.[3] The harmonics in

[1] This is because a slight error in calibration of the heterodyne oscillator causes a large percentage error in the unknown frequency when the latter is small. This error can be minimized by making an initial adjustment of the heterodyne oscillator frequency with a trimmer condenser. This adjustment is made by setting the oscillator frequency to the frequency of the filter (as indicated by output from the fixed amplifier) when the main oscillator dial is set to analyze an unknown wave of zero frequency.

[2] For more detailed information see R. G. McCurdy and P. W. Blye, Electrical Wave Analyzers for Power and Telephone Systems, *Trans. A.I.E.E.*, vol. 48, p. 1167, October, 1929.

[3] This is accomplished by applying a very large voltage to the grid of a highly biased grid. The best tube for the purpose should have two or more

the output wave of such a harmonic generator all pass through a maximum at the same instant as does the fundamental. The relative phase position of a particular harmonic in the wave being analyzed can then be determined by tuning the output of the harmonic generator to exactly maximum response, superimposing the output upon the unknown wave, and shifting the phase of the input to the harmonic generator with a phase-shifting bridge

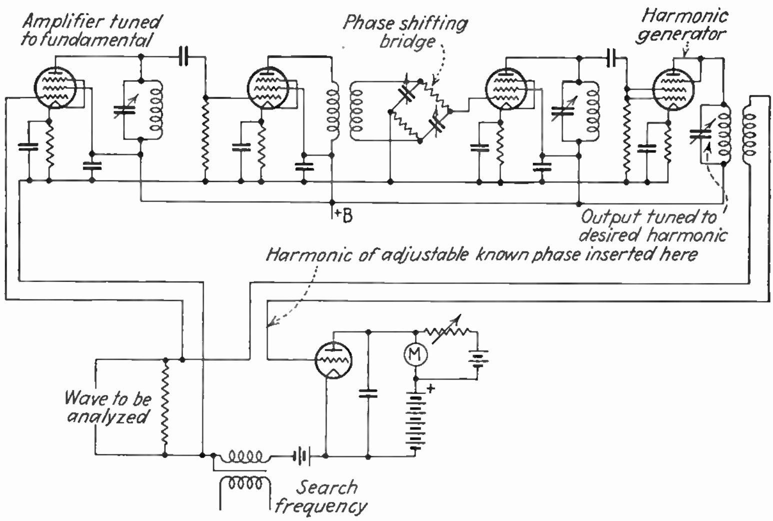



Fig. 88.—Schematic diagram of method for measuring the relative phase of harmonics in a complex wave being measured by the vacuum-tube voltmeter method.

until the harmonic generator is brought in phase with the corresponding harmonic of the unknown wave. The relative phase position of the harmonic in the unknown wave is then determined from the amount of phase shifting that is necessary. The circuit lay-out for applying this method is shown schematically in Fig. 88 for the case of a vacuum-tube voltmeter type of harmonic analyzer.

grids with the two inner grids connected together and acting as the control grid. The grid bias is then preferably obtained by the use of a grid leak. The signal voltage applied to the grid should be as high as possible, preferably 100 volts or more, and the grid-leak resistance should be large, since the greater the leak resistance the smaller the angle of flow. For further information on the factors affecting the angle of flow see F. E. Terman and J. H. Ferns, The Calculation of Class C Amplifier and Harmonic Generator Performance of Screen-grid and Similar Tubes, *Proc. I.R.E.*, vol. 22, p. 359, March, 1934.

**39. Analysis of Radio-frequency Waves.**—Wave form at frequencies up to about 100,000 kc can be observed with the aid of the cathode-ray tube as described in Chap. XIII, and the results may be analyzed by use of a schedule method when desired. The method is not useful in determining very small harmonic components, however, since the width of the cathode-ray spot is great enough to obscure harmonics less than about 5 per cent.

Radio-frequency waves are analyzed experimentally by the substitution method. The fundamental principles involved are illustrated in Fig. 89, where the method is used to determine the harmonic currents flowing in an antenna circuit. A highly

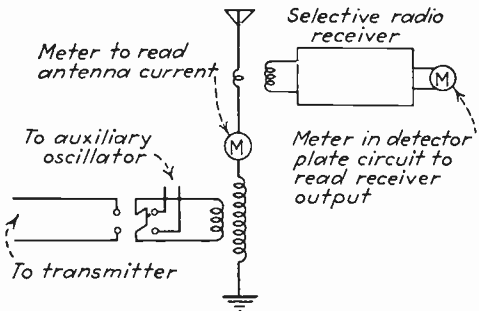



FIG. 89.—Diagram illustrating the measurement of harmonics in the antenna circuit by the substitution method.

selective, properly shielded radio receiver is coupled to the antenna system, tuned to the desired harmonic frequency, and the gain adjusted until a reasonable deflection is observed in the detector plate-circuit meter. The radio transmitter is now turned off and current from an auxiliary oscillator having the same frequency as the harmonic is passed through the coil. The amplitude of the auxiliary oscillator current which gives the original output indication in the radio receiver is then equal to the amplitude of the unknown current. The auxiliary oscillator current can be readily measured since it is not combined with currents of other frequencies.

The amplitude of the harmonics encountered at radio frequencies is usually very small because radio circuits are usually resonant. This places very severe requirements upon the selectivity and cross-modulation characteristics of the radio receiver because of the very large fundamental frequency component that is present, and which must be rejected by the radio receiver. The radio receiver must hence have an adequate number of tuned circuits, and is preferably of the superheterodyne type. The shielding must also be so complete that energy can enter the receiver only at the normal input point, and the selectivity ahead of the first tube must be sufficient to prevent overloading by the large fundamental component, with consequent cross-modulation.

Radio-frequency waves may also be analyzed experimentally by the heterodyne method. The idea is to reduce the frequency of the component being analyzed to some low predetermined frequency, such as 1000 cycles, by use of the heterodyne principle. The 1000-cycle resultant is then amplified in a tuned amplifier and indicated on a suitable output meter. The process of analyzing the wave then consists in superimposing, upon the unknown wave, oscillations from an auxiliary oscillator which is adjusted to give the desired beat note with the frequency component to be measured. The gain of the fixed amplifier is set to give a convenient output, after which the unknown wave is replaced by an adjustable known voltage of the same frequency, and the amplitude adjusted until the standard output is again obtained.[1]

The phase relations in radio-frequency waves are usually unimportant, but if desired can be determined by the same methods outlined for audio frequencies. When it is desirable to determine the relative phase of two radio-frequency voltages of the same frequency, one can employ a cathode-ray tube by applying one voltage to one pair of plates, the other voltage to the other pair, and observing the resulting pattern (see Chap. XIII). Another method of measuring relative phase consists in heterodyning both radio-frequency waves by the same oscillation. In this way it is possible to reduce both waves to audio-frequency waves, which *maintain the same relative phase difference as before the heterodyning process.* The phase differences can then be measured at the audio frequency.[2]

[1] For further details of this method see A. G. Landeen, Analyzer for Complex Electric Waves, *Bell System Tech. Jour.*, vol. 6, p. 230, April, 1927.

[2] The necessary details for carrying out the measurements are given by R. R. Law, A New Radio-frequency Phase Meter, *Rev. Sci. Inst.*, vol. 4 p. 537, October, 1933.

# CHAPTER VII

## VACUUM-TUBE CHARACTERISTICS

**40. Vacuum-tube Coefficients.**—The important coefficients of a triode tube are the amplification factor $\mu$, the plate resistance $R_p$ (sometimes called dynamic plate resistance), and the mutual conductance $G_m$. These can be defined by the following mathematical expressions:

$$\text{Amplification factor} = \mu \quad = \left.\frac{dE_p}{dE_g}\right|_{I_p \text{ constant}} \tag{41a}$$

$$\text{Plate resistance} = R_p = \left.\frac{dE_p}{dI_p}\right|_{E_g \text{ constant}} \tag{41b}$$

$$\text{Mutual conductance} = G_m = \left.\frac{dI_p}{dE_g}\right|_{E_p \text{ constant}} \tag{41c}$$

The exact meaning and significance of these coefficients can be brought out by redefining them in words. Thus the amplification factor represents the ratio of a change in plate voltage to the change in grid voltage which is required to neutralize the effect of the plate-voltage change on the plate current. The plate resistance is the resistance which the plate circuit offers to a small voltage superimposed upon the normal electrode voltage of the tube, *i.e.*, it is the resistance offered to a small alternating-current voltage superimposed upon the plate potential. The mutual conductance is the change in plate current that results per 1-volt change in grid potential.

In screen-grid and pentode amplifiers the tube behavior can be expressed in terms of an amplification factor, plate resistance, and mutual conductance (often called transconductance when applied to tubes with more than three electrodes) which are defined exactly the same as in the case of the triode tube. The actual numerical values of the tube coefficients will depend upon the potentials of the other electrodes, however, and may be greatly different from those encountered in triodes.

In screen-grid and pentode tubes several additional coefficients of some practical importance are introduced by the presence of the screen electrode. The most important of these is $\mu_{sg}$, which can be defined as the relative importance of the control grid as compared with the screen grid in influencing the total space current of the tube. Expressed mathematically, this definition is

$$\mu_{sg} = \frac{dE_{sg}}{dE_g}\bigg|_{\text{total space current and } E_p \text{ constant}} \tag{42a}$$

where $E_{sg}$ and $E_g$ are screen-grid and control-grid voltages, respectively. This coefficient is the one which determines the cut-off point and is for all practical purposes the same as the amplification factor obtained by connecting screen and plate together and using the tube as a triode. It is discussed extensively elsewhere.[1] In screen-grid tubes there is also the factor $\mu_p$ which represents the relative effectiveness of the control-grid and plate potentials in influencing the total space current. This is not identical with the amplification factor $\mu$ which is used to calculate the behavior of an amplifier, as is apparent from the following mathematical definition:

$$\mu_p = \frac{dE_p}{dE_g}\bigg|_{\text{total space current and } E_{sg} \text{ constant}} \tag{42b}$$

Note that in the definition of the amplification factor $\mu$ it is the plate current rather than the total space current that is constant. It is occasionally also necessary to make use of the dynamic resistance of the screen-grid circuit, *i.e.*, the resistance of the screen circuit to a small increment of voltage.

The foregoing coefficients are the only ones which are normally encountered in practical engineering work, although a great many others can be defined. For example, when a triode is operated with the control grid positive, one could conceivably be interested in the dynamic grid resistance, in the reflex amplification factor, *i.e.*, the relative effects which the grid and plate voltages have upon the grid current, and the reflex mutual conductance, *i.e.*, the change of grid current per volt change in plate potential. With screen-grid tubes, pentodes, and hexodes, the number of coefficients which may be defined becomes extremely great since every electrode has an amplification factor with respect to the current of every electrode and also with

[1] See F. E. Terman, "Radio Engineering," Chap. IX, McGraw-Hill Book Company, Inc., 1932.

respect to the total space current, every electrode has a transconductance (or mutual conductance) with respect to every other electrode, and every electrode circuit has a dynamic resistance. With tubes as ordinarily used, most of these coefficients have little if any practical importance, however.

Detection and cross-talk coefficients depend upon second- and third-order coefficients of the tube curves, respectively. The cross-talk coefficient is discussed in Sec. 51, where methods of measuring it are described. The detection coefficients are defined differently by different writers, and because of this it is necessary to consider the detection coefficients in connection with detector theory.[1]

**41. Determination of Vacuum-tube Coefficients from Static Curves.**—The vacuum-tube coefficients defined in the preceding section can be deduced by evaluating the derivatives of the mathematical definitions in terms of small increments taken from the characteristic curves of the tube. This method is not particularly accurate but is always available for rough determinations and gives a clear visualization of what each coefficient represents.

The procedure for determining tube coefficients from the static curves can be made clear by considering some typical examples. One first defines the desired coefficient in terms of small increments. Thus Eqs. (41) become

$$\mu = \left.\frac{\Delta E_p}{\Delta E_g}\right| I_p \text{ constant}$$

$$R_p = \left.\frac{\Delta E_p}{\Delta I_p}\right| E_g \text{ constant}$$

$$G_m = \left.\frac{\Delta I_p}{\Delta E_g}\right| E_p \text{ constant}$$

One can now use the static curve to evaluate the increments called for. Thus if the amplification factor of a triode is to be determined, one must determine from the static curves the ratio of plate- to grid-voltage increments required to keep the plate current constant. This can be done with either $E_g - I_p$ or

[1] A complete treatment of detector theory is given in Chap. VIII of "Radio Engineering." Also see E. L. Chaffee, "Theory of Thermionic Vacuum Tubes," Chaps. XIX, XX, XXI, and XXII, McGraw-Hill Book Company, Inc., 1933.

$E_p - I_p$ curves as shown in Fig. 90. It will be noted that one increment represents the voltage difference between adjacent curves, whereas the other increment is scaled directly off of the coordinate system. The plate resistance is by definition the

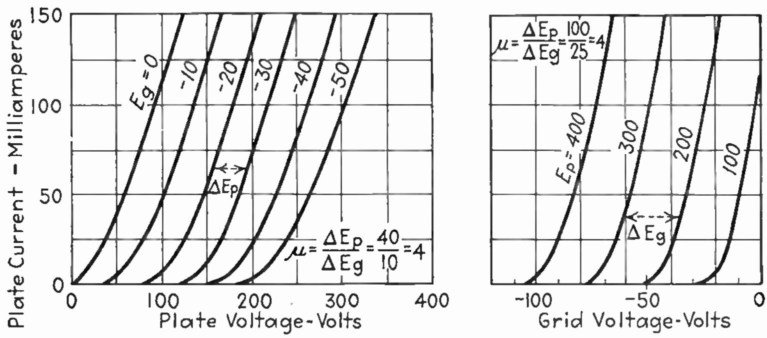



FIG. 90.—Evaluation of triode amplification factor using small but finite increments obtained from curves of tube characteristics. In each case one increment is the potential difference between adjacent curves while the other increment is scaled off as shown.

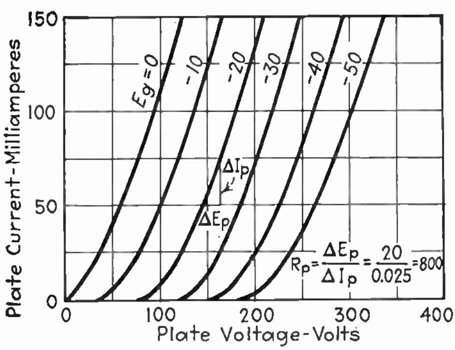



FIG. 91.—Evaluation of triode plate resistance from curves of tube characteristics. Plate resistance is by definition the reciprocal of the slope of the $E_p - I_p$ curve for constant grid bias $E_g$.

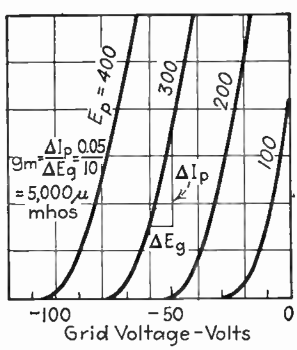



FIG. 92.—Evaluation of triode mutual conductance from the characteristic curves of tubes. The mutual conductance is by definition the slope of the $E_g - I_p$ curve for constant plate voltage.

reciprocal of the slope of $E_p - I_p$ curve for constant grid bias, and so is readily evaluated as shown in Fig. 91. Mutual conductance is likewise by definition seen to be the slope of the $E_g - I_p$ curve for constant plate voltage and so is obtained as shown in

Fig. 92. Plate resistance and mutual conductance can likewise be evaluated from $E_g - I_p$ and $E_p - I_p$ curves, respectively, by methods similar to those employed in evaluating amplification factor.

The coefficients of screen-grid and pentode tubes can be determined in like manner from the characteristic curves. The examples in Fig. 93, showing the determination of plate resistance

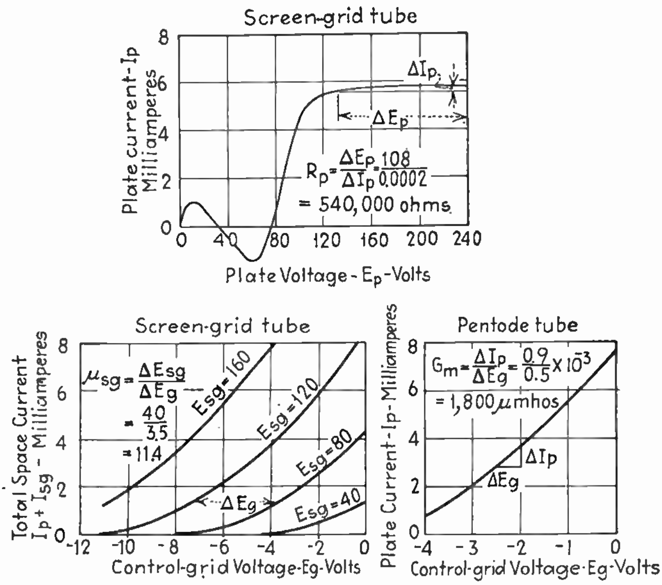



FIG. 93.—Examples showing how coefficients of pentode and screen-grid tubes may be evaluated from characteristic curves of the tubes.

of a screen-grid tube, the mutual conductance of a pentode, and $\mu_{sg}$ of a screen-grid tube, make the method clear. The same constants can be determined from characteristic curves drawn in other ways, and the remaining possible coefficients of tubes with more than three electrodes can be worked out by using the same technique.

In order to obtain reasonable accuracy the characteristic curves must be very accurately drawn on a large sheet of paper, and the increments used must be small. Practically it is difficult to obtain accuracies greater than about 10 per cent, with still greater error to be expected when the coefficients have extreme values such as plate resistances in the order of megohms, amplification factors of 1000, or mutual conductances of a few micromhos.

A modification of the foregoing method of determining tube coefficients, which avoids the necessity of drawing the characteristic curves of the tube, consists in measuring the necessary increments directly with meters. This is sometimes called the incremental method of measuring tube coefficients, and is often found to be a convenient method of making rough determinations. To determine the amplification factor by this method one would add a convenient increment to the grid bias, say $-2$ volts, and then change the plate voltage until the original current was restored. If the plate-voltage change required in this particular instance was 19 volts, the amplification factor would be $^{19}\!/_2 = 9\frac{1}{2}$. The plate resistance can be measured similarly by adding an increment to the plate voltage and reading the resulting increment of plate current, while the mutual conductance is obtained by adding an increment to the grid bias and noting the change of plate current while the plate voltage is maintained constant. The practical difficulty with the incremental method is that if the increments are small they cannot be read accurately unless rather complicated circuit arrangements such as shown in Fig. 94 are employed to read the increments directly, while if the increments are large the accuracy is poor because the exact definitions given in Sec. 40 call for very small increments.

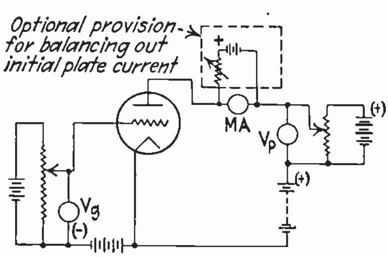



FIG. 94.—Circuit for accurate measurement of increments to plate and grid voltages and to plate current, for the purpose of determining coefficients.

**42. Dynamic Methods of Measuring Tube Coefficients—Bridge Circuits.**—The most accurate method of determining tube coefficients involves the use of small alternating current and voltage increments, combined with a circuit arrangement that permits evaluation of the coefficients by a null balance in a pair of telephone receivers.

Bridge circuits of this type in which the balance conditions depend upon the tube coefficients are shown in Fig. 95, together with the equations of balance.[1] In the circuits for measuring plate resistance, the plate circuit of the tube is made the "unknown" arm of an ordinary alternating-current bridge, so that the results give the resistance of the plate circuit to the alternating current which the bridge superimposes upon the steady electrode potential. The circuits for measuring mutual conductance apply to the grid of the tube a small alternating voltage developed by the voltage drop in resistance $R_1$, and balance out the resulting alternating plate current by resistances $R_2$ and $R_3$. The amplification factor is similarly measured by applying to the grid the alternating voltage developed across the resistance $R_1$, and neutralizing the effect which this has upon the plate current by means of the voltage developed across resistance $R_2$.

*Reactive Balance.*—In all of the arrangements shown in Fig. 95 the capacities associated with the circuit give rise to reactive currents that will obscure the proper balance point unless prevented from flowing through the telephone receivers. These capacities are shown in Fig. 95 for the case in which the cathode of the tube is at ground potential, which is the natural point for

[1] The derivation of these equations follows:

*Plate Resistance.*—These circuits are simple Wheatstone-bridge circuits with the plate resistance to be evaluated placed in the $X$ or *Unknown* branch.

*Mutual Conductance.*—The current $I_0$ produces a voltage drop $I_0R_1$ that is applied to the grid and causes an alternating plate current $G_mI_0R_1$ to flow. In circuit $c$ this current is balanced out of the telephone receiver by an equal and opposite current through $R_2$ produced by the voltage $I_0R_3$ developed across $R_3$ (assuming $R_2 >> R_3$ so that the shunting effect of $R_2$ on $R_3$ can be neglected). Hence $G_mI_0R_1 = I_0R_3/R_2$ and solving for $G_m$ gives the condition of balance for circuit $c$. It is to be noted that the impedance of the telephone receivers has no affect on the results.

In the case of circuit $d$ the alternating plate current produces a voltage drop $G_mI_0R_1R_2$ across $R_2$, and this is balanced by the drop $I_0R_3$ in $R_3$, so that $G_mI_0R_1R_2 = I_0R_3$. Solving for $G_m$ gives the balance equation. It will be observed that in this circuit the mutual conductance as measured is equal to $\mu/(R_p + R_2)$ and so is equal to the true mutual conductance $\mu/R_p$ only if $R_p >> R_2$.

*Amplification Factor.*—Here the voltage applied to the grid in each case is $I_0R_1$, and the effect of this on the plate current is balanced by a voltage $I_0R_2$ developed across $R_2$. The amplification factor is then by definition the ratio of these; *i.e.*,

$$\mu = \frac{I_0R_2}{I_0R_1} = \frac{R_2}{R_1}$$

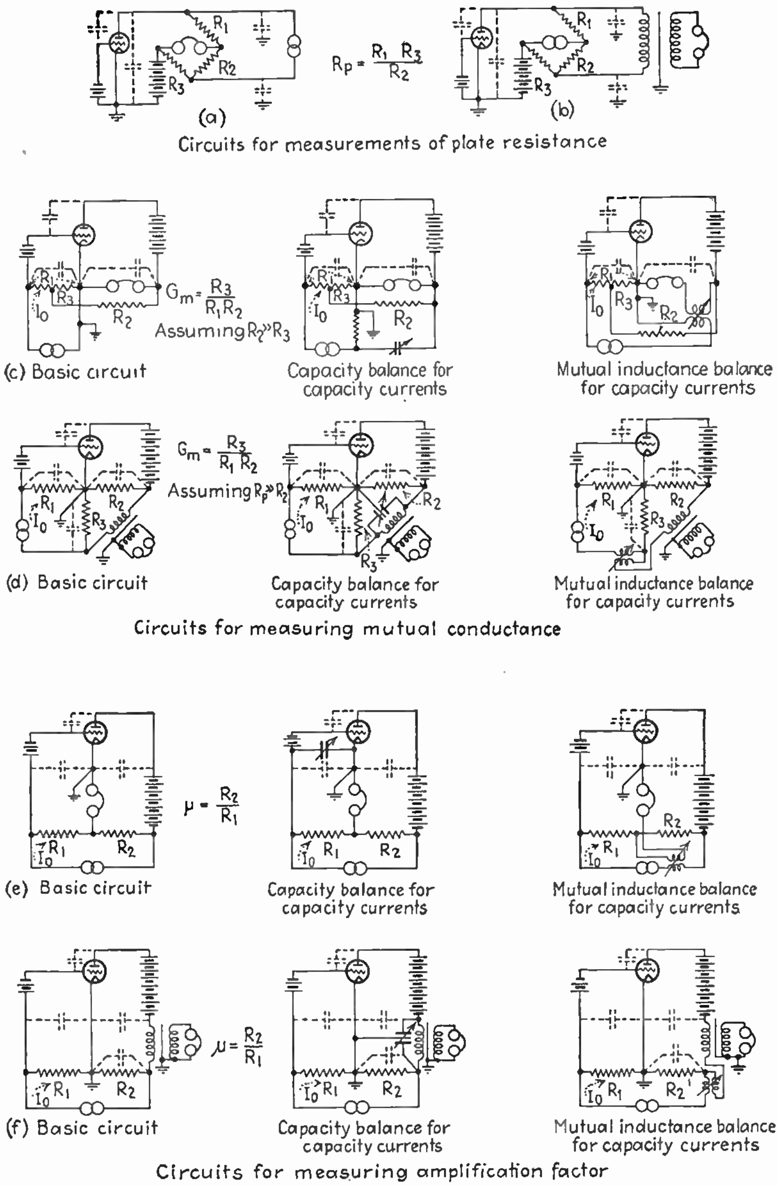



Fig. 95.—Bridge circuits for measuring coefficients of triodes, together with the equations of balance, and means for balancing out reactive currents.

grounding ordinary vacuum-tube circuits. Means of balancing the reactive currents out of the telephone receivers are shown in the same figure. In the two circuits for measuring plate resistance it will be observed that all the capacities are in shunt with either the $X$ or the $R_3$ arms, and so can be taken care of by a variable condenser across one of these arms to provide a power-factor balance. In the circuit shown at $c$ for measuring mutual conductance, the only important capacity current passing through the telephones is that which flows through the grid-plate tube capacity, and this can be neutralized by a variable condenser in shunt with $R_2$. In circuit $e$, the capacities in shunt with $R_1$ and $R_2$ when considered in relation to these resistances are seen to form a Wheatstone bridge which can be brought into balance by a variable condenser from grid to cathode. Circuits $d$ and $f$ require that the effective capacity from the two sides of the output transformer to ground be the same, and this can be most conveniently realized by the use of a double stator condenser as shown in the figures.

It is also possible to balance out reactive currents by the insertion of a small quadrature voltage in series with the telephone receivers as shown on the right-hand side of Figs. 95*c* to 95*f*, inclusive. This is accomplished by placing the primary coil of a variometer in series with the oscillator leads, and connecting the secondary in series with the telephone receivers (or the primary of the output transformer).

*Miscellaneous Considerations.*—The two circuits given for measuring each of the tube coefficients differ primarily in the location of the telephone receiver and the oscillator, and are of equal merit except that circuit $e$ cannot be used when the control grid draws current. Under this rather unusual operating condition, the quantity measured is

$$\left.\frac{dE_p}{dE_g}\right|_{I_p + I_g \text{ constant}}$$

whereas the amplification factor is defined as

$$\left.\frac{dE_p}{dE_g}\right|_{I_p \text{ constant}}$$

In every measuring arrangement where the telephone receivers are not at ground potential, Fig. 95 shows an output transformer having a grounded shield between primary and secondary. This transformer is necessary under such circumstances, since otherwise capacity currents will flow from the telephone receivers to their wearer, and considerable hand or body effect will be present.

A number of considerations affect the choice of the resistance arms of the bridge circuits. In the first place, the magnitudes

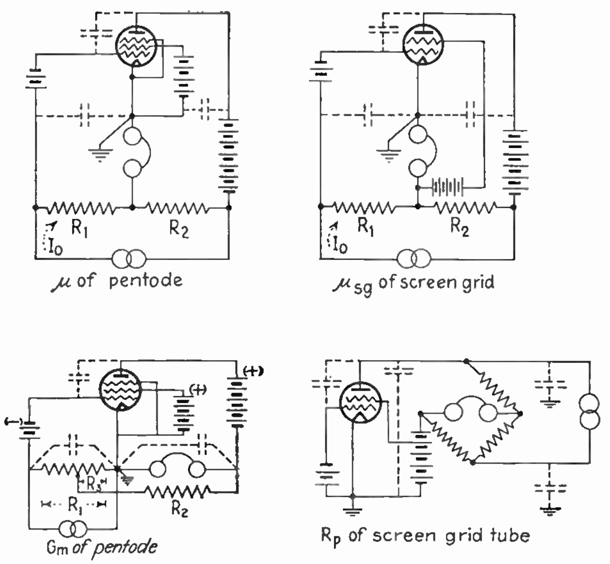



FIG. 96.—Methods of applying the bridge circuits of Fig. 95 to the determination of pentode and screen-grid tube coefficients.

of the resistances should be so chosen in relation to the resistances of telephone receivers, oscillator circuit, and transformer primaries that the direct-current voltage drop will be a minimum, and will also be as nearly constant as possible as the bridge is being balanced. The resistances must also be capable of carrying the plate currents to be encountered, which sometimes exceed the safe limits of commercial general-purpose bridges and resistance boxes. When amplification factor is measured, it is customary to make $R_1$ a fixed resistance of 10 ohms in ordinary cases, and 1 ohm when the amplification factor is high, while $R_2$ is a variable resistance, often a decade resistance box. Common circuit proportions for measuring mutual conductance by circuit $c$ are $R_1 = 50$ ohms, $R_2 = 20{,}000$ ohms, and $R_3$ variable, while for circuit $d$ one can use $R_1 = 1000$ ohms, $R_3 = 100$ ohms, and $R_3$ variable. In measuring mutual conductance by circuit $c$ the shunting effect of $R_2$ on $R_3$ makes it necessary to multiply the measured results by the factor $R_2/(R_3 + R_2)$ if great accuracy is required. Likewise in circuit $d$ it will be noted that the measurements actually give $\mu/(R_p + R_2)$, so that the true mutual conductance is obtained by multiplying the measured results by $(R_p + R_2)/R_p$. It will be noted that by proper choice of circuit proportions these correction factors can be made negligibly small in any particular case.

*Coefficients of Screen-grid and Pentode Tubes.*—All the discussion given so far for the bridge measurement of tube coefficients has been specifically applied to triode tubes, but the same principles may obviously be applied to screen-grid and pentode tubes with minor modifications. Typical examples of the circuits as applied to such tubes are shown in Fig. 96. The dynamic resistance of any electrode circuit is measured by making this circuit the unknown arm of a Wheatstone bridge, just as before. Likewise the mutual conductance between any two electrodes is determined by applying the drop in the resistance $R_1$ to one electrode and balancing out the resulting current that flows in the circuit of the other electrode by the resistances $R_2$ and $R_3$. The various amplification factors $\mu$, $\mu_{sg}$, $\mu_p$, etc., are all determined by applying the voltage drop developed across $R_1$ to one electrode, balancing the resulting effect produced in the other electrode circuit by a potential developed across $R_2$, and locating the telephone receivers in the part of the circuit where the current is to be constant. The methods for balancing the reactive current in the telephone receivers are the same as in the case of triodes.

Special bridge circuits can be devised for the measurement of negative resistance, negative mutual conductance, negative amplification factors. Bridge circuits for measuring negative resistance are discussed in considerable detail in Sec. 21, and circuits for measuring these other special cases are described in the literature,[1] but are of too little practical importance to warrant the space required for further consideration here.

The oscillator indicated in Figs. 95 and 96 should have a frequency of the order of 1000 cycles and should not apply over 1 or 2 volts to the tube electrodes in order that the coefficients as measured may approach those defined in Sec. 40 in terms of derivatives. It is desirable that the oscillator be coupled into the bridge circuit through an input transformer located close to the bridge, and having a grounded electrostatic shield between primary and secondary windings. This keeps down the stray capacities introduced by the oscillator and also distributes them more evenly. Some type of input transformer is absolutely essential if the oscillator has one terminal at ground potential, as is the case in some commercial equipment. If accurate results are to be obtained, the sources of power which supply plate, grid, screen, etc., potentials must be thoroughly by-passed to the oscillator frequency, since otherwise the resistance of the power supply would be included as part of the resistance of the tube circuits.

The bridge circuits of Figs. 95 and 96 are widely used in temporary set-ups for measuring tube coefficients, and when properly handled are capable of the same accuracy as other bridge circuits. There are, however, certain important defects which limit their usefulness. Thus an appreciable direct-current voltage drop is introduced in the circuits, as already mentioned; the bridge used in measuring electrode resistance is at a high direct-current potential to ground; while in the circuits for measuring mutual conductance and amplification factor the batteries supplying electrode voltages must be ungrounded and have a low capacity to ground, thus practically preventing the use of rectifier-filter systems.

**43. Dynamic Methods of Measuring Tube Coefficients—Voltage-ratio Method.**[2]—In this method the tube coefficients are evaluated in terms of the ratio of two voltages injected into the tube circuits in such a way that the ratio required to give a null balance is a measure of the desired coefficient. A third voltage is normally provided for balancing out

[1] See E. L. Chaffee, "Theory of Thermionic Tubes," Chap. IX, McGraw-Hill Book Company, Inc., 1933.

[2] For further information on this method see W. N. Tuttle, Dynamic Measurement of Electron Tube Coefficients, *Proc. I.R.E.*, vol. 21, p. 844, June, 1933.

reactive currents in the telephone receivers arising from stray capacities.

The application of the voltage-ratio method to the measurement of the various coefficients of tubes is shown schematically in Fig. 97, together with the equations for balance.[1] It will be observed that the amplification factor is obtained by balancing the effect of the voltage $e_1$ applied to the grid by the voltage $e_2$ acting in the plate circuit. The dynamic plate resistance is determined by injecting a voltage $e_2$ in the circuit and balancing the resulting current that flows by means of voltage $e_1$ and $R_s$. The mutual conductance is determined by applying the voltage $e_1$ to the electrode that is serving as the control, and balancing the resulting current that flows into the output electrode by the use of $e_2$ and $R_s$.

The voltages $e_1$ and $e_2$ are normally derived from a common source and have their ratio controlled by means of attenuators located between the source and the point where the voltage is developed. In order to obtain accurate results, it is necessary that the internal resistance of the sources of $e_1$ and $e_2$ as viewed by the tube be low compared with the dynamic resistance of the tube circuits, the resistance $R_s$, and the reactance of the various stray capacities around the circuit. This is necessary in order that $e_1$ and $e_2$ may not depart appreciably from their open-circuit values when injected into the tube circuits. An internal resistance of the order of 10 ohms is satisfactory for ordinary conditions.

A variety of circuit arrangements can be used to produce the voltages $e_1$, $e_2$, and $e_3$. A typical example, showing the

[1] These equations are readily derived by writing down the relation that must exist in the equivalent circuit shown in Fig. 97 for no current in the telephone receivers. For amplification factor,

$$\mu e_1 = e_2 \qquad \text{or} \qquad \mu = \frac{e_2}{e_1}$$

For mutual conductance,

$$\frac{\mu e_1}{R_p} = \frac{e_2}{R_s} \qquad \text{or} \qquad \frac{\mu}{R_p} = G_m = \frac{e_2}{e_1}\frac{1}{R_s}$$

For plate resistance,

$$\frac{e_2}{R_p} = \frac{e_1}{R_s} \qquad \text{or} \qquad R_p = \frac{e_2}{e_1}R_s$$

It will be noted that the impedance of the telephone receivers does not enter into the result.

numerical values suitable for ordinary receiving-tube measurements is shown in Fig. 98 and indicates what is required. The essential features of the arrangement in Fig. 98 include a shielded input transformer with separate well-insulated secondaries for $e_1$, $e_2$, and $e_3$, thus maintaining these voltages accurately in the same phase and same ratio with respect to each other, and at the same time providing isolation to direct-current voltages. The magni-

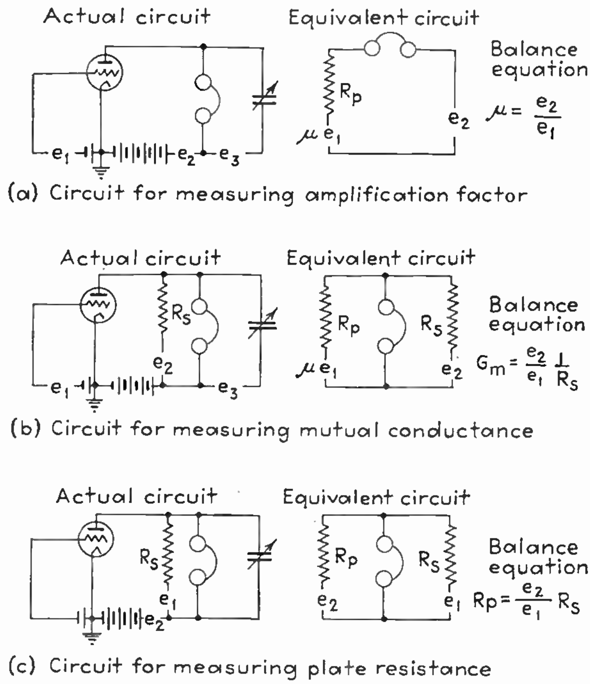
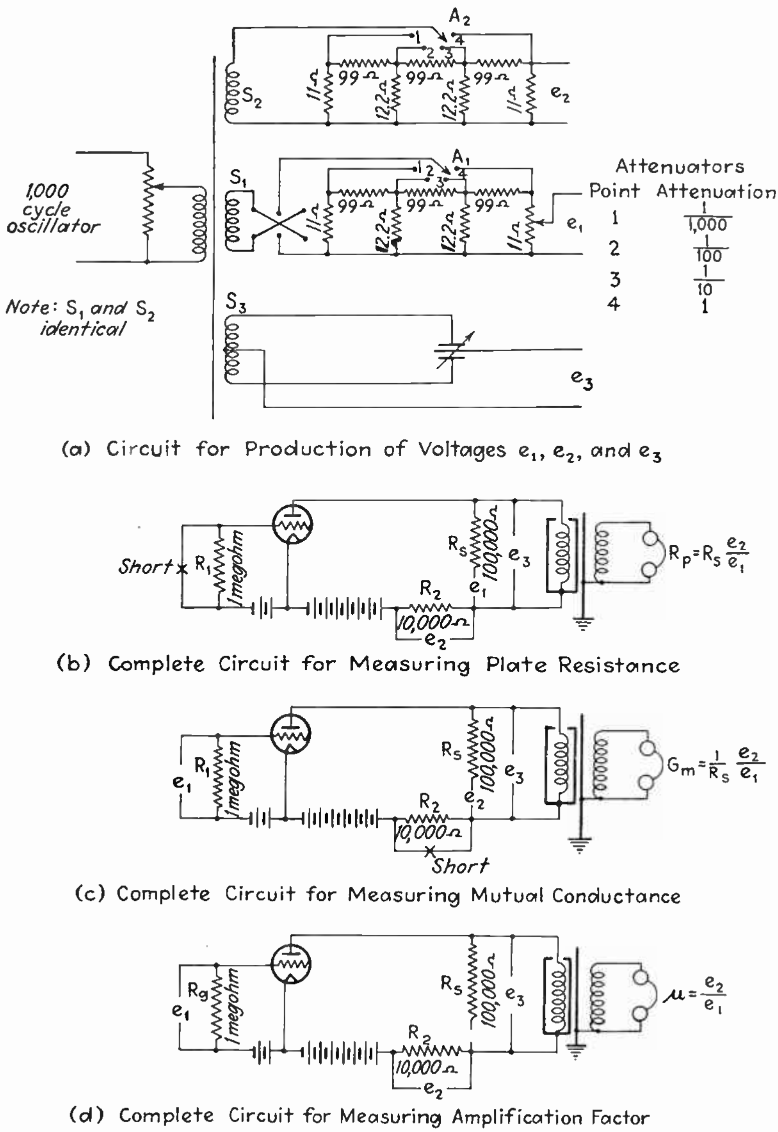



FIG. 97.—Schematic diagram illustrating the voltage-ratio method of measuring tube coefficients.

tude of $e_2$ is controlled by attenuator $A_2$ in which each stage of attenuation reduces the voltage by a factor of 10. The magnitude of $e_1$ is controlled by an attenuator $A_1$ similar to $A_2$, together with a tapered slide-wire potentiometer $P$ for giving a continuous variation between attenuator steps.

The theoretical range of tube coefficients which can be measured with the circuit proportions shown in Fig. 98, assuming that the variable attenuator will control the voltage over a 10 to 1 variation, is




Note: $R_1$ *is for the purpose of preventing grid circuit from opening momentarily during switching and is shorted when not shunted by* $e_1$. $R_2$ *serves a similar purpose and is shorted when not shunted by* $e_2$.
*Both* $R_1$ *and* $R_2$ *are high compared with output resistance of* $e_1$ *and* $e_2$, *so have negligible effect on these voltages*

FIG. 98.—Circuit diagram showing how the voltages required for measuring tube coefficients by the voltage-ratio method may be produced.

Amplification factors from 10,000 to 0.001.
Plate resistances from 1000 megohms to 100 ohms.
Mutual conductances from 100,000 micromhos to 0.01 micromho.

Practically, some of these extreme values cannot be determined with accuracy. This is particularly true of the dynamic resistance, since dielectric losses in stray capacities introduce appreciable error when resistances much above 1 megohm are measured.

Care must be taken in the laying out of the circuits shown in Fig. 98 because of the wide difference in power levels which may exist between various parts of the circuit. Proper spacing is essential between parts which may operate at widely different power levels, as, for example, the input and output of an attenuator system or the outputs of the two attenuators. It is particularly important that no coupling exist between the input and output transformers. This can be achieved by employing the maximum possible space while at the same time orientating the two for minimum mutual inductance. The output transformer must have primary and secondary windings carefully shielded, since the primary winding is not at ground potential in certain of the measurements. The primary winding of the output transformer should have a very low direct-current resistance in order to carry the normal plate current, but should at the same time have an inductance of at least 1 henry.

Stray couplings existing in the equipment can be discovered by making suitable checks. Magnetic coupling to the output transformer can be detected by adjusting the circuits to give a balance, and then rotating one of the transformers by hand. If there is no coupling whatsoever, the position of the transformer has no effect. The effectiveness of the electrostatic shielding of the output transformer can be determined by attempting to measure the plate resistance of a tube with a cold filament. The resulting value should be in excess of 10 to 100 megohms, and, if not, indicates excessive dielectric losses in the circuit, something wrong with the shielding system, or that $e_1$, $e_2$, and $e_3$ are not in the same phase. Another check upon the shielding and phase relations is to arrange the connections as one would in measuring amplification factor, but with the tube filament cold. It should then be possible, by adjusting the capacity balance, to obtain a balance in the telephone receivers even when $e_2$ is large.

Correct phase relations between $e_1$ and $e_2$ can be insured by making the secondary windings $s_1$ and $s_2$ identical and symmetrical with respect to the primary and operating them into identical resistances. This does not insure that $e_3$ has the proper phase, however, and trouble may be experienced from the fact that winding $s_3$ operates on an open circuit, and thus may deliver an output voltage $e_3$ having a phase differing from that of the voltages $e_1$ and $e_2$. Any phase error of this type can be detected by measuring the plate resistance of a tube with the filament cold. If the resistance measured is absurdly low, and if the resistance varies greatly when a low-loss capacity is shunted from plate to filament, phase errors are present, and must be corrected by inserting a phase shift between $s_3$ and $e_3$.

The different coefficients can be determined by switching the voltages $e_1$, $e_2$, and $e_3$, together with the telephone receivers and the capacity balance, into the circuit at the appropriate places. A multi-pole tap switch can be wired up to make the appropriate connections for measuring the more important coefficients such as $\mu$, $R_p$, and $G_m$, and also possibly certain of the special coefficients involved with screen-grid and pentode tubes. One position of the switch should connect the various voltages to binding posts, so that special connections can be set up for measuring unusual coefficients. In laying out a switch arrangement, care must be taken to see that damage cannot result when the various tube circuits are opened. This is particularly important in the case of the grid, since in some instances the loss of the grid bias will seriously damage the tube. To avoid this possibility it is usually desirable that the grid and plate circuits always be closed through a high resistance as shown in Fig. 98. In laying out the switching arrangement it will also be noted that the entire assembly of $R_s$, output transformer, and capacity balance $e_3$ can be switched as a unit.

The measuring equipment can be checked by measuring known resistances. These can be placed across the plate-to-filament terminals of the tube socket. After this has been done, a final check consists in determining the amplification factor, mutual conductance, and plate resistance of an actual tube and noting whether or not $G_m = \mu/R_p$. Discrepancies should not exceed a few per cent.

Any seldom-used coefficient not discussed above can be obtained by introducing the voltage $e_1$ and $e_2$ and the telephone receivers at suitable places in the circuit. Negative coefficients can be readily measured by merely reversing the leads to either $e_1$ or $e_2$.

The voltage-ratio method of measuring tube coefficients is particularly suitable in permanent laboratory set-ups for deter-

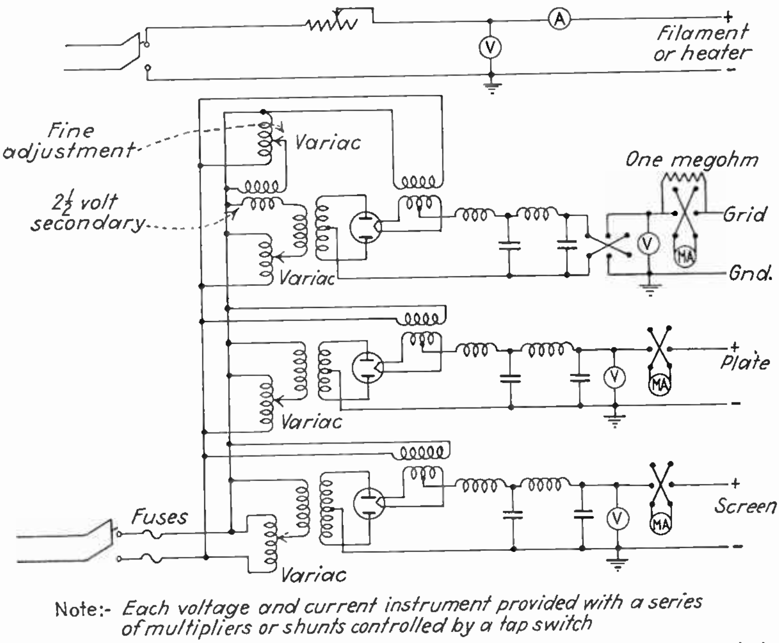



FIG. 99.—Power-supply system suitable for determining the characteristic curves of receiving and small power tubes.

mining tube characteristics. This method overcomes all of the important limitations of the bridge circuit, since the batteries may be operated with one terminal grounded and the direct-current drops in the measuring equipment can be made so low as to be negligible under ordinary circumstances. The resulting equipment is extremely flexible, being capable of measuring almost every imaginable tube coefficient of almost any conceivable value that could be encountered. Furthermore, the measuring procedure for determining the different coefficients is substantially the same, and the results are direct reading except for the decimal point. The accuracy is entirely satisfactory for all ordinary requirements, being usually within 1 or 2 per cent with the equipment shown in Fig. 98, and can be made equal to that of any bridge circuit by refinements of detail.

**44. Apparatus for Measuring Tube Characteristics.**—In laboratories where it is frequently necessary to run characteristic curves on tubes, it is desirable to provide permanent equipment for this purpose. The details of such equipment may be varied considerably, but a general idea of what is suitable for handling receiving tubes and small power tubes is shown in Fig. 99, which gives the circuit diagram of equipment that has been found to be particularly satisfactory in actual use. The arrangement shown in Fig. 99 develops plate, screen-grid, and control-grid voltages from rectifier-filter systems while employing a storage battery for filament power. The equipment provides for control of the various voltages and includes meters for reading the voltage and current at each electrode, with a sufficient number of multipliers to make it possible to measure with reasonable accuracy any value that may be desired. Good voltage regulation of the various power-supply units is obtained by the use of a Variac to control the voltages, supplemented by a fine adjustment for the control-grid voltage. Good voltage regulation can be insured by using low-impedance rectifier tubes (either 5Z3 or hot-cathode mercury-vapor types), together with a choke-input type of filter system designed to draw sufficient bleeder current to prevent the voltage from soaring at no load. All meter ranges must be properly fused to prevent damage as a result of improper manipulation of the controls or as a result of defective tubes. Reversing switches must be provided for plate, screen, and control-grid currents, and a reversing switch for the grid voltage is also essential. In order that a negative grid bias may not be lost as a result of a burned-out fuse, the meter in the grid line should be shunted with a 1-megohm resistor as shown.

The multiplicity of tube types creates a problem which is apparently most easily solved by permanently wiring up one group of sockets to handle the more common tube types, while at the same time bringing out the various potentials to a series of binding posts. One can then provide a panel containing one socket of each of the various standard sizes and types with all of the No. 1 pins of these sockets wired together and brought out to one binding post, all of the No. 2 pins similarly wired together and brought out to a second binding post, and so on. By patching between the binding posts representing the various pins and the binding posts at which the various electrode voltages appear, it is possible to set up any socket combination that may be desired.

The equipment as described above makes it a simple matter to determine the characteristic curves of tubes. By combining this equipment with a unit for measuring the tube coefficients by the voltage-ratio method described above, it is then possible to set the tube for any desired operating condition and obtain the resulting coefficients with a minimum of effort.

Equipment such as described for obtaining tube characteristics is conveniently mounted on a relay rack with the tube sockets at the top, the apparatus for determining tube coefficients on a panel just below the sockets, below which will be panels containing meters and multipliers, voltage-control equipment, and finally power-supply units, together with the switches and monitoring lights. To make the equipment foolproof, it is desirable to provide lights on the main switches, a light which comes on if the grid reversing switch is in the positive grid-potential position, and a light which comes on if any of the main voltage-control units are not set at zero. These last lights must not turn off when the main switch is thrown to *off*.

Large power tubes can be tested with equipment following the same general principles but modified as occasion demands. Where very large amounts of power are to be handled, the voltage control is normally obtained with the aid of a grid-controlled rectifier (mercury-arc, thyratron, or grid-glow tube) or by the use of a saturated-core reactor in series with the transformer primaries. In the saturated-core method a thyratron or similar type of grid-controlled rectifier is used to vary the direct-current saturation of the core, and hence the voltage that appears across the transformer primary.[1]

**45. Tube Checkers, Mutual Conductance Meters, Etc.**—A variety of tube-testing devices have been developed for use in

[1] Examples of thyratron-controlled power-supply systems are given by H. W. Lord, A Life Test Power Supply Utilizing Thyratron Rectifiers, *Proc. I.R.E.*, vol. 21, p. 1097, August, 1933; and C. B. Foos, A Vacuum Tube Controlled Rectifier, *Elec. Eng.*, vol. 53, p. 568, April, 1934.

connection with the servicing of radio receivers. These are normally rather crude devices as compared with the testing equipment described above, but are entirely satisfactory for the purpose of detecting defective tubes.

Most of these devices operate by checking the emission of the cathode in some manner. The simplest way of doing this is to apply an alternating voltage to the plate and read the resulting direct-current component of the plate current. The grid may be left at zero potential, may have a negative bias developed by a battery, or may have a small alternating voltage applied to it with such a polarity as to make the grid negative at the same instant that the plate is positive. Some test equipment provides

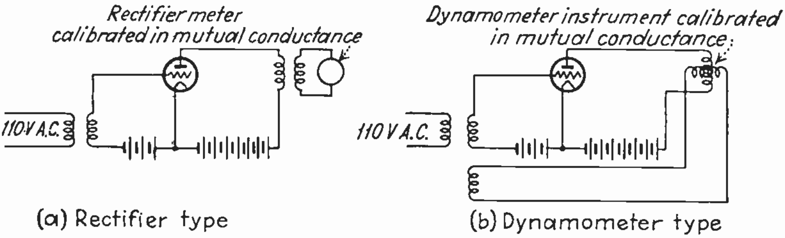



FIG. 100.—Circuit arrangements of two types of mutual-conductance meters. The diagrams do not show the provision which is normally made for adjusting the voltage on the transformer to a standard value.

means for changing the grid bias when a button is depressed. The change of plate current that results as the bias is changed in this way is indicative of the mutual conductance of the tube. Pentode, screen-grid, and similar tubes can be tested in such equipment by connecting the extra electrodes to the plate and operating the tube as a triode. The most elaborate outfits for checking tubes have a small rectifier-filter system, and attempt to approximate the conditions under which the tube is normally operated. The plate current that flows, and the change of plate current with a small change in grid bias, is then a certain indication of the condition of the tube.

The simplest means of accurately checking the condition of a tube is by determining the mutual conductance, and several simple arrangements have been devised for indicating mutual conductance directly upon the dial of a small meter. Two possible arrangements are illustrated in Fig. 100. In both of these a known alternating voltage derived from the 60-cycle power system through a step-down transformer is applied to the grid, and the resulting alternating component of plate current is measured by an indicating instrument which is calibrated directly in mutual conductance. In the arrangement shown at Fig. 100*a* the alternating component of plate current is measured by coupling a small rectifier instrument to the plate circuit of the tube through a transformer, while in the circuit of Fig. 100*b* the plate current is passed directly through one coil of a dynamometer instrument and a known 60-cycle current is passed through the other coil. The deflection is then proportional to the alternating component of plate current, and hence to the mutual conductance. It will be noted that in both these arrangements the direct-current component of plate current has no effect upon the instrument deflection. The accuracy of both methods depends upon the constancy of the 60-cycle voltage, and provision not shown in Fig. 100 is usually made for adjusting the voltage on the transformer to a standard value by reconnecting the meter so that it serves temporarily as a voltmeter.

# CHAPTER VIII

## AUDIO-FREQUENCY AMPLIFICATION

**46. The Measurement of Amplification.**—The most important characteristics of audio-frequency amplifiers are the amount of amplification obtainable and the way in which this amplification varies with frequency. These quantities are measured either by determining the voltage that must be applied to the amplifier input at different frequencies to maintain a constant

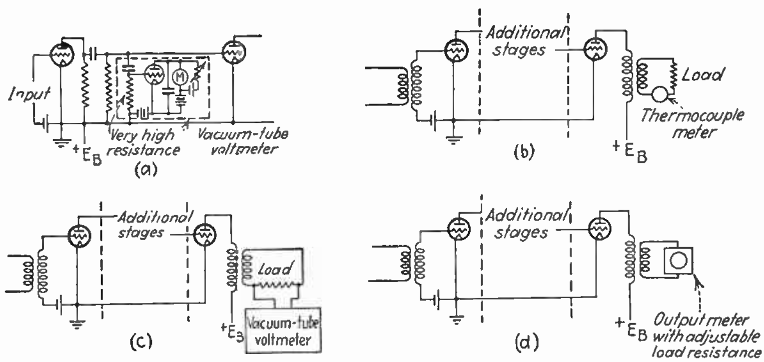



Fig. 101.—Methods of measuring output voltage of audio-frequency amplifiers.

known output, or by measuring the output that is obtained when a constant known voltage of varying frequency is applied to the amplifier input. The former method is usually preferred because it avoids all possibility of overloading the amplifier.

The output level of a power stage can be determined as shown in Figs. 101*b*, 101*c*, and 101*d* by measuring the current in the load resistance with a thermocouple instrument, determining the voltage across the load with a vacuum-tube voltmeter, or using an output meter such as described in Sec. 9 to supply the proper load resistance and measure the power. No special precautions are necessary because the output is developed across a comparatively low impedance.

The output voltage of a stage of voltage amplification is usually measured with a vacuum-tube voltmeter as shown in Fig. 101*a*, and care must be taken to keep the input capacity of the vacuum-tube voltmeter as small as possible so that the effect upon the amplification characteristic will be small. This shunting capacity of the measuring equipment can be kept below 10 $\mu\mu$f by keeping the grid lead of the voltmeter as short and direct as possible, and by using in the vacuum-tube voltmeter a type of tube having a very low input capacity.

In making amplification measurements upon individual stages of a multistage amplifier it is always absolutely necessary that the stage in question operate into its normal load, and that all subsequent stages of amplification be in operation.

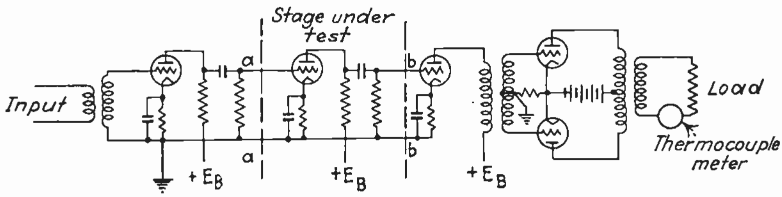



FIG. 102.—Arrangement for testing performance of single stage of multistage amplifier.

This is because the input impedance of the grid into which the individual stage operates has considerable effect on the amplification characteristics, and also because the subsequent stages may be the cause of regeneration that will alter the amplification of the stage being investigated. The most accurate method of determining the characteristics of an individual stage of a multistage amplifier, such as stage *ab* in Fig. 102, is first to apply the test voltage across the input to the stage under investigation (across *aa* in Fig. 102) and measure the amplification from this point to the output. The test voltage is then applied across the output of the stage being studied (across *bb* in Fig. 102) and the amplification between this point and the load obtained. The ratio of these two amplifications is obviously the amplification of the stage between points *aa* and *bb*. The result corresponds to actual operation since the stage operates into its normal load impedance, is subject to the same regenerative action with respect to the stages of higher power level as in actual operation, and no shunt impedances that change the amplification characteristic are used in making the measurements. The only thing altered is the regeneration between the stage under test and stages of lower power level, and this has negligible effect on the result.

*Production of Known Audio-frequency Voltages.*—Known audio-frequency voltages from about 1 mv upward can be produced by applying a known voltage to a calibrated voltage divider as shown in Fig. 103*a*, or by passing a known current through a known resistance as in Fig. 103*b*. The voltages $E$ indicated in Fig. 103 are those obtained on open circuit, and have an equivalent internal impedance $R_{eq}$ as shown in the figure. When the voltage is applied to a load as at Fig. 103*c* rather than an

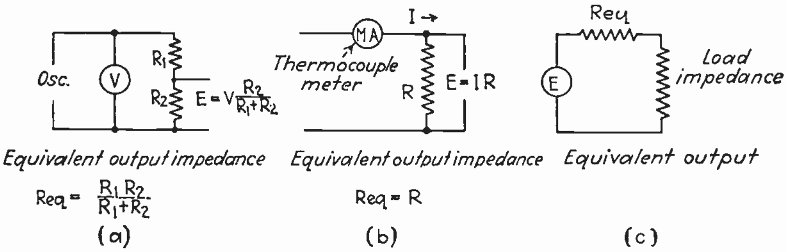



FIG. 103.—Simple methods of producing known voltages, together with equivalent output circuit for calculating reduction in voltage produced by a load.

open circuit, this internal resistance must be considered as being in series with the load impedance, with the open-circuit voltage applied to the combination as shown.

When audio-frequency voltages of the order of 1 mv or less are required, or when it is necessary to cover a wide range of magnitudes, it is more satisfactory to employ some form of adjustable attenuating network as described in Sec. 23. The requirements which such an attenuator should meet are constant output impedance and continuous variation of output voltage. A great many suitable arrangements are possible, and several of these are shown in Fig. 104. The circuit shown at Fig. 104*a* employs a number of T attenuating networks which can be switched in and out of the attenuator as desired, combined with a thermocouple voltmeter across the input which can be used to interpolate between attenuator steps. At Fig. 104*b* the input voltage is maintained at a predetermined constant value, and interpolation between attenuator steps obtained by means of the slide wire, which has a second variable resistance operated from the same shaft to keep the output impedance constant. The circuits at *c* and *d* differ from *a* and *b* in that the T sections have been replaced by a ladder attenuator. The arrangement

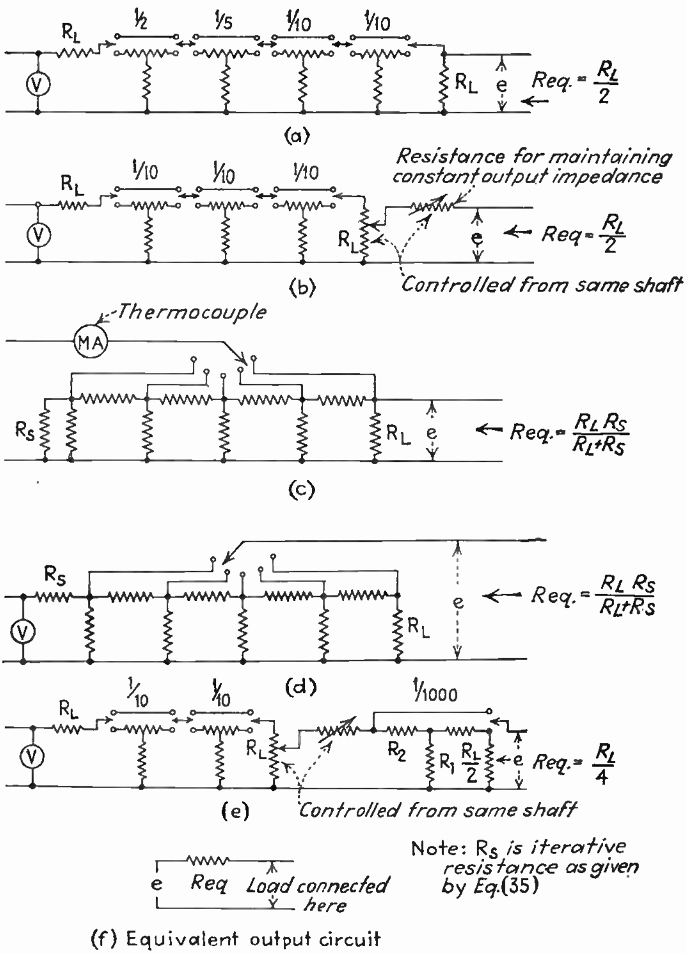



FIG. 104.—Attenuator systems for producing known audio-frequency voltages.

at *e* is sometimes desirable because the slide wire here is not required to operate at extremely low power levels. It will be noted that here the attenuator in the slide-wire output must be designed to operate between resistances equal to half of the resistance, $R_L$, and that the equivalent output impedance is now $R_L/4$ instead of $R_L/2$ as in Fig. 104*b*.

All of the circuits of Fig. 104 have a constant output resistance of the value indicated in the figure. The effect of a load placed across the output terminals can hence be calculated by assuming that the output resistance is in series with the load, and considering that the no-load output voltage of the attenuator acts on the combination as shown at Fig. 104*f*.

*Phase Shift in Amplifiers.*—The phase shift between amplifier input and output voltages is sometimes of importance because it may alter the relative phase of different frequencies being amplified, and because it also represents a time delay that may not be the same for the different frequencies being amplified.

The simplest method of measuring phase shift in amplifiers is by means of a cathode-ray tube. The output voltage is applied to one pair of plates of the tube, while a voltage having the same phase as the input voltage is applied to the second pair. The phase difference between the two voltages can then be calculated from the character of the pattern that results, as described in Sec. 79. Phase shift can also be accurately determined by bridge arrangements such as shown in Fig. 105. Here the voltage developed at the output point is balanced in both magnitude and phase by an equal and opposite voltage. This is accomplished by applying to the amplifier input a voltage derived by passing a current through a known resistance, while deriving the balancing voltage by passing the same current through an adjustable resistance and an adjustable mutual inductance. The voltage induced in the secondary of the mutual inductance is in quadrature with the voltage developed in the resistance and can be reversed in phase by the switch shown in the figure. The different forms of the circuit shown in Fig. 105 are necessary to cover all possible values of amplification and the full range of phase shifts from 0 to 360.° The equations for balance in each case are given in the figure. Balance can be determined by the use of telephone receivers operating from a shielded output transformer.

*Use of Decibels to Express Relative Amplification.*—The variation of amplification with frequency in audio-frequency amplifiers is often expressed in decibels referred to some arbitrary level which is taken as zero decibels. The significance of such curves can be understood by considering what a decibel means. The decibel is a unit for expressing a power ratio and is given by the relation

$$\text{Decibels} = \text{db} = 10 \log_{10} \frac{P_2}{P_1} \tag{43}$$

The decibel has no other significance; and if it is to be used in expressing relative amplification, it therefore signifies power

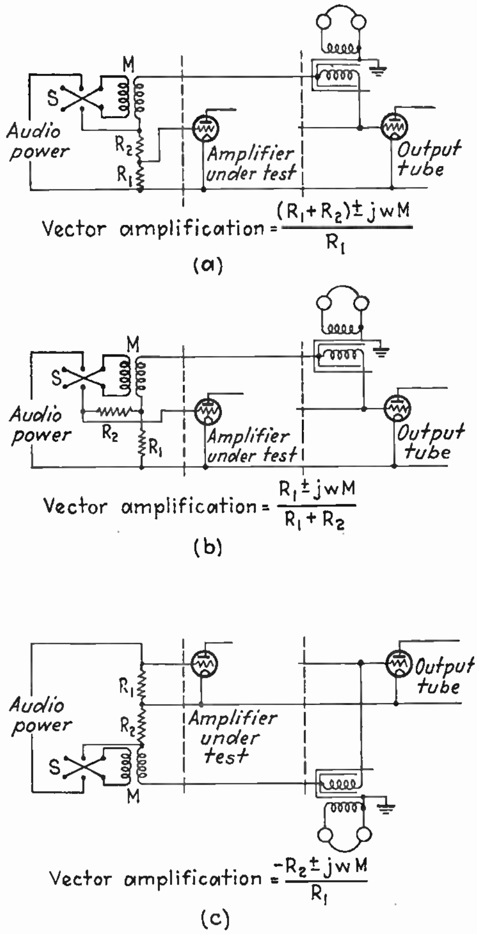



FIG. 105.—Circuits for determining phase shift in audio-frequency amplifiers.

output as a function of frequency with respect to some arbitrary power. Thus if the output voltage varies with frequency as shown in Fig. 106*a*, one might replot this curve in decibels by assuming some arbitrary power as the standard. This might, for instance, be the power output obtained at 400 cycles. The power output at any other frequency is then proportional to $(E/E_{400})^2$ where $E$ is the voltage output at the frequency in question and $E_{400}$ is the output voltage at 400 cycles. Since the power output under these conditions is proportional to the

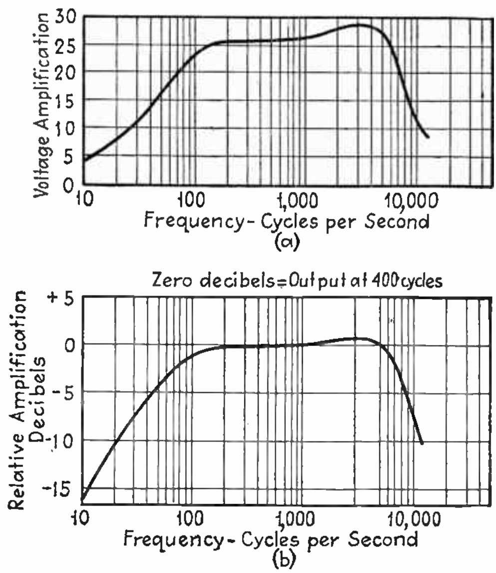



FIG. 106.—Illustration of how relative amplification can be expressed in decibels.

square of the voltage, one can rewrite Eq. (43) as follows for this particular case:

$$\text{db} = 10 \log_{10} \frac{P_2}{P_1} = 10 \log_{10} \left(\frac{E}{E_{400}}\right)^2 = 20 \log_{10} \left(\frac{E}{E_{400}}\right) \quad (44)$$

It is now possible to plot a curve giving amplification in terms of decibels as is done at Fig. 106*b*. The significance of the decibel curve can be seen by considering a specific case. Thus the fact that the amplification in Fig. 104*b* is 5 db lower at 45 cycles than at 400 cycles means that the output power at 45 cycles is 0.316 times the power at 400 cycles.

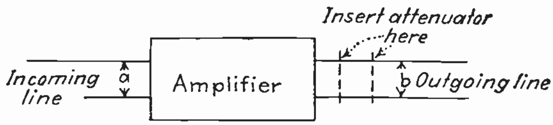
From the foregoing it is seen that anything which increases or decreases the amplification can have its effect expressed in terms of decibels. Thus if one introduces an extra stage of amplification which increases the output voltage twenty times, then the gain in output power is $(20)^2$ or 400 times, and this power ratio when substituted in Eq. (43) is seen to represent a power gain of 26 db.

It will be noted the decibel is fundamentally a power unit. It cannot be used to express voltage ratios except in so far as these voltages are related to power ratios. If two voltages are applied to identical resistances, then the resulting powers are, of course, proportional to the square of the voltages; but if the voltages are applied to different resistances, then it is necessary to take into account this fact if the decibel unit is to be employed.

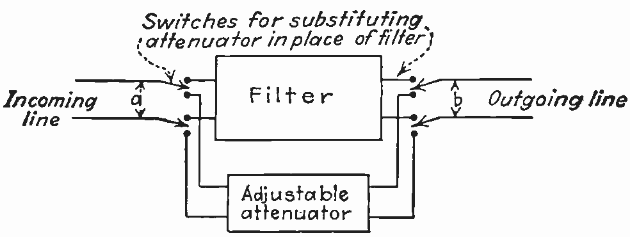
Power levels in audio-frequency amplifiers are often expressed in decibels referred to a standard level of 6 milliwatts. Thus an output transformer rated for service at +36 db level is capable of handling a power output that is 36 db above 6 milliwatts, or 24 watts. Likewise, a microphone rated at −50 db will deliver an output power 50 db less than 6 milliwatts, or $6 \times 10^{-8}$ watts when spoken into under average conditions. If the microphone has a 200-ohm internal impedance, this power will be developed in a load resistance of 200 ohms and represents an r.m.s. voltage of $(6 \times 10^{-8} \times 200)^2 = 0.0035$ volt, which can then be increased to the desired level by the use of an amplifier provided with a step-up input transformer to match the grid of the first tube to the 200-ohm microphone.

**47. Measurements of Power Gain and Power Loss.**—In certain circumstances, particularly in telephone work, it is desirable to determine the effect which equipment in the circuit has upon the power delivered to the circuit output. Thus when an amplifier is inserted in a long telephone line, one is interested in how much increase in power results. Again, if a filter is inserted in an electrical circuit, it is quite important to know the effect which the insertion of the filter has on the power level at the load.

Power gain or loss under such circumstances can be measured by the same technique as that described in connection with audio-frequency amplifiers, but it is often more convenient to employ a substitution method such as illustrated in Fig. 107. At Fig. 107*a* the problem is to determine the power level at the point $b$ when the amplifier is inserted in the circuit as compared with the power level with the amplifier removed. In the absence of the amplifier the power levels at points $a$ and $b$ are obviously the same, while with the amplifier the power level at $b$ will naturally be greater than at $a$. If, however, one inserts an attenuator as indicated in the figure, and this attenuator is designed to match the line impedance, then the power level at point $b$ can be made equal to the power level at point $a$




(a) Substitution method of measuring amplifier gain




(b) Substitution method of measuring insertion loss of filter

FIG. 107.—Typical circuit arrangements for measuring power gain and loss by the substitution method.

by introducing enough attenuation to equal exactly the gain of the amplifier. It is thus possible to evaluate the amplifier gain directly in terms of the attenuation which must be introduced to neutralize the gain.

The same procedure is modified in Fig. 107$b$ to determine the loss caused by the insertion of a filter into the circuit. With the filter present, the power level at point $b$ will be less than at point $a$ as a result of losses in the filter. If the filter is now removed and replaced by an attenuator which has the same input and output impedances as the lines, then by adjusting the attenuator until the power levels at $a$ and $b$ are the same as when the filter is present, the insertion loss of the filter is naturally equal to the loss introduced by the attenuator.

The chief advantage of the substitution method of measuring power gain and loss is that the accuracy of the measurement depends only upon the accuracy of the attenuator, which being made of resistance elements is extremely stable. One does not depend in any way upon the accuracy of the instruments used to determine the power level at points *a* and *b*, since it is merely necessary that these power levels have the same value when the attenuator is present as when it is absent.

**48. Amplitude Distortion and Load Limit in Amplifiers.**—The load-carrying capacity of an amplifier is limited by the fact that, as the input voltage to the amplifier is increased, the wave form of the output voltage departs more and more from the input wave form. The principal consequence of this action is that an output wave contains frequencies not present in the signal being amplified. The distortion that results is commonly measured in terms of the r.m.s. amplitude of the harmonics expressed as a percentage of the fundamental frequency component of the output wave. When the signal being amplified is small, the percentage distortion defined in this way is virtually zero, but as the signal is increased the percentage of harmonics first increases slowly and then very rapidly. The load limit of an amplifier is defined as the point at which the percentage of harmonics has some arbitrary value, usually 5 per cent.

The amount of distortion present in an amplifier can be determined qualitatively by observing the wave shape of the amplifier output in either a cathode-ray or magnetic-type oscillograph. It is possible in this way to detect distortions as small as 5 per cent, and from the manner in which waves are distorted it is usually possible to determine the cause of the distortion, as illustrated in Fig. 108. Quantitative analysis of distortion can be made by photographing these wave forms and determining the frequency components by use of the schedule method described in Sec. 37.

Distortion may be determined experimentally by analyzing the output wave with one of the harmonic analyzers described in Sec. 38. When this is done, it is essential that the input impedance of the analyzer either have the same value for all frequencies, so that it may be considered as part of the load, or have negligible effect upon the load. To have negligible effect when analyzing a current wave means that the input impedance of the analyzer must be small compared with the impedance of the circuit into which the analyzer is introduced, while in analyzing voltage waves the analyzer input impedance must be high compared with the impedance across which it is connected. These impedance requirements arise from the fact that distortion in amplifiers depends upon the load impedance offered the fundamental and the harmonics.

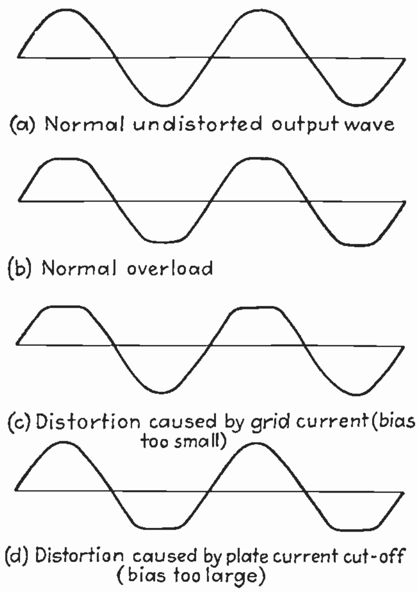



FIG. 108.—Oscillograms of output current of a Class A power amplifier tube showing how the cause of distortion may be determined from the wave slopes.

Since the load impedance of an amplifier affects the amount of distortion, and since the load impedance will inevitably vary somewhat with frequency, it is necessary to make the distortion measurements at different frequencies if a complete picture of the amplifier behavior is to be obtained. For many purposes it is sufficient, however, to make distortion measurements at a single representative frequency such as 400 cycles. Simplified equipment can then be employed since fixed filters will separate the harmonics and permit their measurement.[1]

[1] See A. W. Barber, A Simplified Harmonic Analyzer, *Electronics*, vol. 1, p. 374, November, 1930.

A pure sine-wave input is essential in making distortion measurements. In order to insure this, it is usually necessary to employ a suitable filter, since the output of laboratory oscillators often contains from 5 to 15 per cent harmonics.

*Analysis of Output Waves of Vacuum Tubes from the Dynamic Characteristic.*—When the load into which a vacuum tube operates is a resistance, it is possible to work out the dynamic characteristic according to conventional methods.[1] From this dynamic characteristic one can plot the output wave point by point as shown in Fig. 109, and analyze the result by the schedule method outlined in Sec. 37. This analysis can be greatly simplified by taking the origin at the crest of the wave, since then the curve is symmetrical about the origin and hence contains no sine terms when expressed by Eq. (39*b*). The schedule method then reduces to the formation of a few simple sums and differences, provided only the first few harmonics are required. Thus the coefficients $A_0$, $A_1$, and $A_2$ (*i.e.*, the *change* of the d-c plate current, the fundamental, and second harmonic, respectively) can be determined from a knowledge of the instantaneous plate current when the applied voltage is at zero, maximum, and minimum, according to the formulas

$$\text{D-c component} = A_0 = \frac{I_{\max} + I_{\min} - 2I_0}{4} \tag{45a}$$

$$\text{Fundamental} = A_1 = \frac{I_{\max} - I_{\min}}{2} \tag{45b}$$

$$\text{Second harmonic} = A_2 = \frac{I_{\max} + I_{\min} - 2I_0}{4} \tag{45c}$$

$$\frac{\text{Second harmonic}}{\text{Fundamental}} = \frac{A_2}{A_1} = \frac{I_{\max} + I_{\min} - 2I_0}{2(I_{\max} - I_{\min})} \tag{45d}$$

The notation is indicated in Fig. 109.

When third- and fourth-harmonic components can be expected in the output, as will be the case with pentodes and with badly overloaded triodes, one can determine the components by obtaining from the dynamic characteristic the instantaneous plate

[1] See, for example, the author's book "Radio Engineering," p. 159. The dynamic characteristic of push-pull Class A amplifiers can be worked out by extending this method as described by B. J. Thompson, Graphical Determination of Performance of Push-pull Audio Amplifiers, *Proc. I.R.E.*, vol. 21, p. 591, April, 1933.

current when the applied voltage wave is at zero, maximum, and minimum, and at 0.707 of the maximum and minimum values.

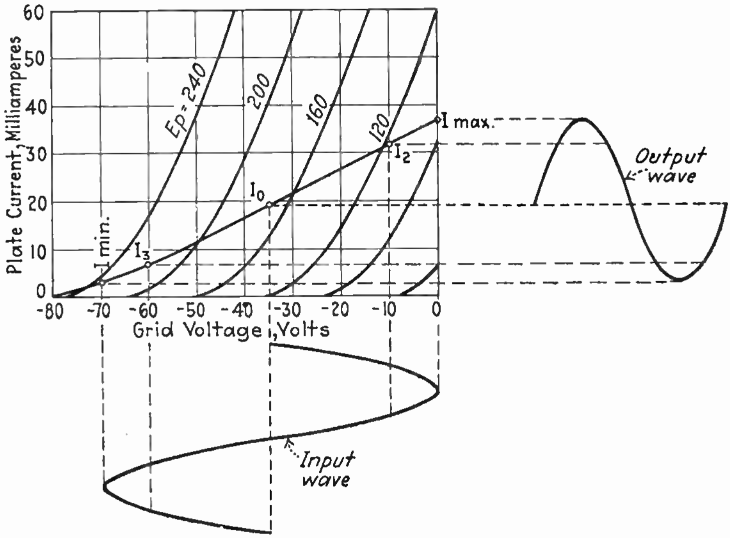



FIG. 109.—Typical dynamic characteristic for a resistance load, showing how the output wave shape can be derived.

The formulas then are[1]

$$\text{D-c component} = A_0 = \frac{\frac{1}{2}(I_{\max} + I_{\min}) + I_2 + I_3 + I_0}{4} \tag{46a}$$

$$\text{Fundamental} = A_1 = \frac{\sqrt{2}(I_2 - I_3) + I_{\max} - I_{\min}}{4} \tag{46b}$$

$$\text{Second harmonic} = A_2 = \frac{I_{\max} + I_{\min} - 2I_0}{4} \tag{46c}$$

$$\text{Third harmonic} = A_3 = \frac{I_{\max} - I_{\min} - 2A_1}{2} \tag{46d}$$

$$\text{Fourth harmonic} = A_4 = \frac{2A_0 - I_2 - I_3}{2} \tag{46e}$$

The notation is as shown in Fig. 109, where $I_2$ and $I_3$ are currents at the two 0.707 points.

The dynamic-characteristic method of determining distortion has the advantage that it requires no information other than

[1] See G. S. C. Lucas, Distortion in Valve Characteristics, *Exp. Wireless and Wireless Eng.*, vol. 8, p. 595, November, 1931.

the characteristic curves of the tube, but it is limited to resistance loads. Since ordinary loads have some reactance because of the output transformers, etc., the dynamic-characteristic method gives results in practice that are merely good approximations suitable primarily for comparative purposes.

The change in d-c plate current which occurs when the tube is amplifying, and which is given by the coefficient $A_0$ in Eqs. (45*a*) and (46*a*), produces a change in the operating conditions which must be taken into account if the maximum accuracy is to be obtained. The dynamic characteristic as ordinarily worked out assumes that the load offers the same resistance to the d-c plate-current increment $A_0$ that it offers to the various alternating components flowing in the plate circuit, and this has the effect of causing the actual direct-current potential applied to the plate to be reduced by the amount $A_0R_L$ (where $R_L$ is the load resistance) when the tube is amplifying. Actual loads normally offer no impedance to the change in plate current, however, since the load is usually coupled into the plate circuit through a transformer. The result is that a dynamic characteristic for a plate-supply voltage $E_B$ must be worked out as though the voltage was $(E_B + A_0R_L)$ if the highest accuracy is to be obtained.

*Overload Characteristics of Amplifiers.*—The approximate point at which an amplifier overloads in actual operation can be determined by relatively simple measurements of the amplifier behavior. The characteristics of all amplifier tubes are such that, when appreciable overloading occurs, the r.m.s. value of the output voltage (or current) is not proportional to the input voltage but rather increases at a slower rate. This is shown in Fig. 110 where it is seen that the point of departure from the ideal linear relation occurs almost exactly when the harmonic content increases rapidly.[1]

In Class A amplifiers, distortion occurs either as a result of driving the grid positive so that grid current is drawn, or by driving the grid so negative that plate current cut-off is approached. Since driving the grid positive produces grid

[1] An ingenious method for making use of this principle to test amplifiers on a production basis is given by Arthur E. Thiessen, The Accurate Testing of Audio Amplifiers in Production, *Proc. I.R.E.*, vol. 18, p. 231, February, 1930.

current, while driving the grid close to cut-off causes the d-c plate current to become greater than with no signal, the presence of these forms of distortion can be determined by inserting direct-current meters in the plate and grid circuits. In making use of the change of plate current which occurs with distortion, one must remember that the change of plate current is roughly proportional to the amount of distortion present, and that some distortion can always be tolerated.

In a Class A amplifier employing a single output tube, practically the only distortion frequency produced up to the load limit is the second harmonic; but as the load limit is exceeded higher order even harmonics, and also odd harmonics of considerable amplitude, are produced as illustrated in Fig. 110. In a Class A push-pull amplifier the even harmonics are completely canceled if the two tubes have identical characteristics, but the odd harmonics are not affected. A push-pull amplifier, therefore, has practically no distortion up to the load limit, but beyond this the distortion increases rapidly. With pentodes the tube characteristics are such that both even and odd harmonics are present in considerable magnitude at even small outputs, and the percentage of harmonics present at the rated output is usually quite high.

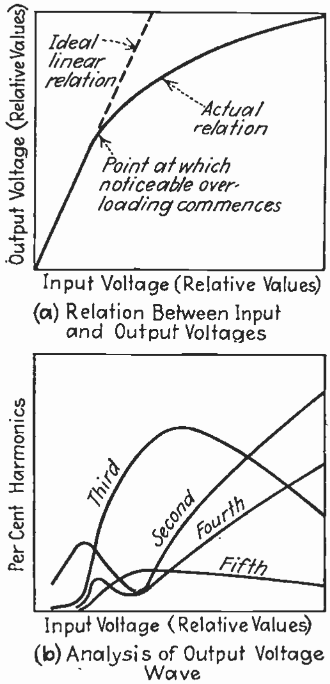



FIG. 110.—Variation of magnitude and distortion of output voltage as the input voltage of a single-tube Class A power amplifier is varied. In a push-pull Class A amplifier the behavior would be almost exactly the same except that the second and fourth harmonics would be absent.

In Class B amplifiers the even harmonics are negligible if the tubes are identical, and the principal harmonic is then the third. Class B amplifiers also develop small amounts of high-order harmonics which are present even when the output is small, and are a result of the commutator action that takes place as the load is transferred between tubes. In making measurements of distortion in Class B amplifiers it is necessary to consider the driver stage as part of the amplifier since a considerable part of the distortion produced is a result of the voltage regulation of the driver output.

*Special Operating Characteristics of Class B Amplifiers.*—Class B audio-frequency amplifiers operate under voltage and current relations that differ greatly from those encountered in Class A

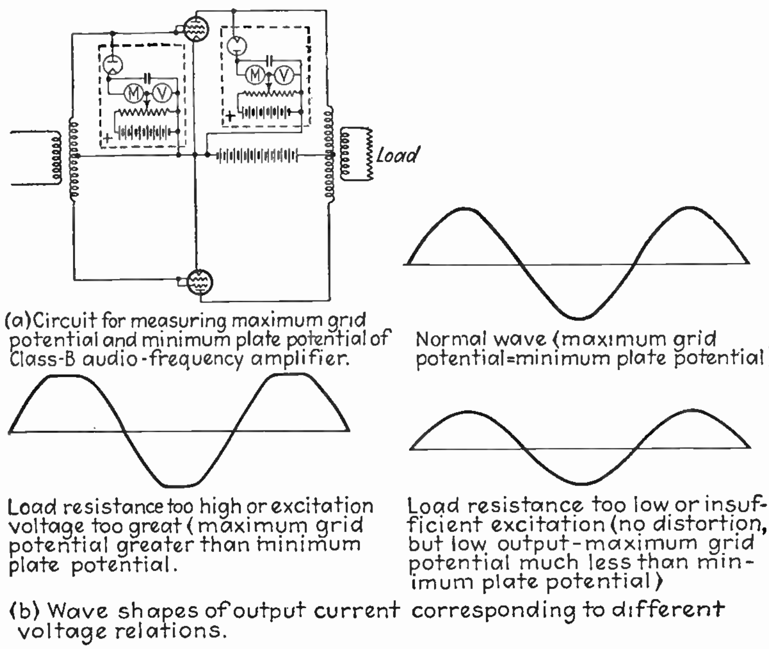



Fig. 111.—Class B amplifier with crest and trough meters for determining essential voltage relations in the tube, together with wave shapes of output current for different voltage relations.

amplifiers. In order to avoid distortion it is necessary that the minimum instantaneous plate potential never become less than the maximum positive potential reached by the grid, for otherwise the grid robs the plate of a disproportionate share of the total space current and the wave of plate current acquires a flat top that introduces distortion as shown in Fig. 111. With a given grid excitation the minimum instantaneous plate potential becomes less as the load resistance in the plate circuit is increased, and at the same time the power output and efficiency both increase. Hence in order to obtain the maximum undistorted power output with a given grid excitation, the load resistance should be as large as possible without making the minimum plate voltage less than the maximum grid potential. If, on the other hand, the load resistance is kept constant and the grid excitation varied, then an excessive excitation has the same effect as an excessive load resistance, *i.e.*, makes the minimum plate potential fall below the maximum grid potential, thus introducing distortion. Likewise insufficient excitation gives the same result as insufficient load resistance, and hence results in low output and low efficiency.

It is possible to determine what is going on in a Class B amplifier by measuring the maximum positive potential reached by the grid with a positive-crest meter such as described in Sec. 7, and determining the minimum instantaneous potential reached by the plate with a trough meter such as is also described in Sec. 7. Circuit arrangements for using these instruments are shown in Fig. 111 and are similar to those shown in Fig. 125 and discussed in Sec. 54 for making similar measurements on Class C power amplifiers.

**49. Audio-frequency Transformer Constants.** *Interstage Audio Transformers.*[1]—Interstage audio transformers are used for coupling the plate of one stage of an amplifier to the grid of the succeeding stage, and are characterized by a step-up voltage ratio which commonly ranges from 2 to 6, and by the fact that the load connected across the secondary is equivalent to a small capacity. When the transformer is used in this way, the complete equivalent circuit taking into account the input and output circuit elements is shown in Fig. 112*b*. In this figure the driving tube has been replaced by an equivalent generator having an internal resistance equal to the plate resistance of the tube, while a capacity equal to the plate-cathode capacity is shunted across the primary. The output tube to which the secondary voltage is applied represents a load equal to the input admittance of this tube.

The equivalent circuit of Fig. 112*b* can for all practical purposes be reduced to unity turn ratio and simplified as shown in Fig. 112*c*, in which the capacity shunting the primary, the hysteresis

[1] Much of the material in this section is based upon a research carried on by I. E. Wood while a graduate student at Stanford University.

loss, the mutual capacity between primary and secondary, and the load resistance $R_g$ have all been omitted.[1] The resistance

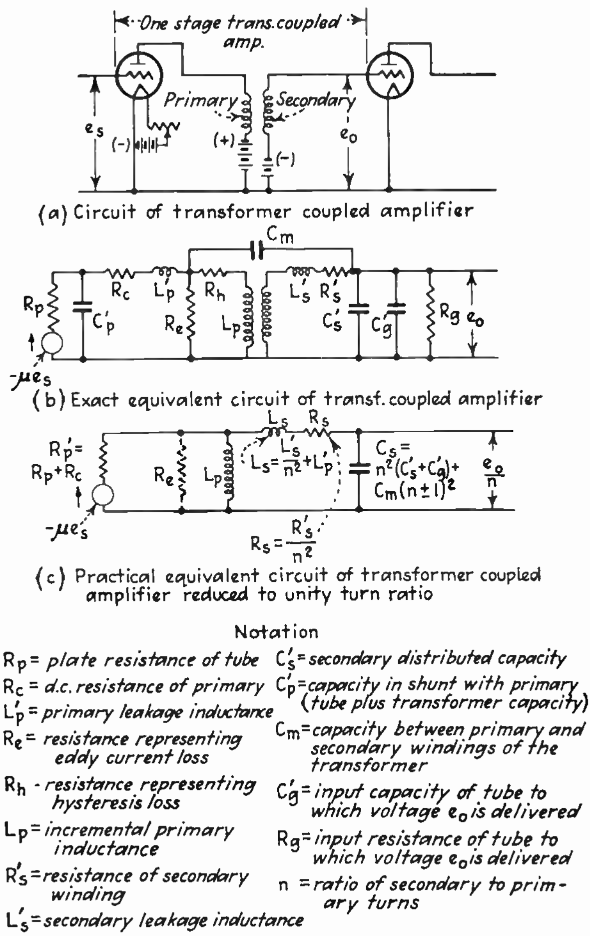



FIG. 112.—Actual circuit of a transformer-coupled amplifier, together with exact and approximate equivalent circuits.

$R_e$ that represents eddy-current losses is shown dotted in Fig. 112c because it can also be neglected unless high accuracy is essential. This resistance is always very large and is independent of fre-

[1] The justification for these various simplifications is that they have negligible effect upon the behavior of the amplifier. Thus the capacities shunting the primary have an extremely high impedance compared with the

quency and core saturation provided magnetic skin effect is absent. The justification for lumping the primary and secondary leakage inductances together is that most of the leakage inductance is in the secondary, and furthermore that at the frequencies where the leakage inductance is of importance the exact division between primary and secondary is of no significance. The important elements of the final simplified circuit of Fig. 112*c* are the step-up ratio $n$ of the transformer, the primary inductance $L_p$, the primary copper resistance $R_c$, which is in effect an addition to the plate resistance of the tube, the leakage inductance $L_s$, and the effective secondary capacity $C_s$. The significance of these various properties and their use in calculating the amplification characteristic are considered in detail elsewhere[1] but in brief it can be said that the amount of amplification is controlled largely by the step-up ratio, that the extent to which the amplification falls off at low frequencies is determined by the primary inductance $L_p$, and that the way in which the amplification varies with frequency at the higher frequencies is controlled by $L_s$ and $C_s$. The frequency at which $L_s$ and $C_s$ are in resonance is particularly important as it represents the approximate high-frequency limit for which uniform amplification is possible.

The primary inductance $L_p$ that is effective in the equivalent circuits of Fig. 112 is the incremental inductance measured with low alternating flux density and the appropriate d-c current, by using technique such as described in Sec. 15.

The eddy-current resistance $R_e$ can usually be ignored. It is to a high degree of accuracy equal to the impedance at the primary terminals when the frequency is such that the capacity $C_s$ and the primary inductance $L_p$ are in parallel resonance. This can be measured with a dynatron bridge of the type described in Fig. 54 or by an ordinary impedance bridge.

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impedance with which they are in shunt, while the mutual capacity is negligible because of the way in which transformers are used with the side of the secondary next to the primary at ground potential. What small mutual capacity exists can be replaced with fair accuracy by assuming that the distributed secondary capacity $C_s'$ is modified slightly. Similarly the resistance $R_g$ is usually too high to be of importance at audio frequencies, while the hysteresis resistance $R_h$ is too small to have much effect.

[1] See the author's book "Radio Engineering," pp. 142–156.

The step-up ratio of the transformer may be most satisfactorily measured by applying a known voltage to the primary and measuring the potential at the secondary terminals with a vacuum-tube voltmeter. This measurement must be carried out at a low frequency, *i.e.*, a few hundred cycles or less, since at higher frequencies incipient resonance between the capacity $C_s$ and the leakage inductance $L_s$ will make the voltage that appears at the secondary terminals higher than the voltage actually induced in the secondary winding. When measurements of step-up ratio at low frequencies are made, the actual voltage acting on the primary inductance is less than the voltage applied to the primary terminals by the drop in the direct-current resistance of the primary winding. The error thus introduced can be allowed for by multiplying the apparent step-up ratio by the factor $\sqrt{1 + (R_c/\omega L_p)^2}$ where $R_c$ and $L_p$ are primary resistance and inductance, respectively.

The leakage inductance $L_s$ reduced to unity turn ratio can be readily determined by measuring the inductance between the primary terminals when the secondary is short-circuited. The leakage inductance is independent of frequency and of saturation effects in the core, since the leakage paths are in air, so the measurement may be made at any frequency and without passing direct current through the primary. It is not permissible to measure leakage inductance from the secondary terminals by short-circuiting the primary since errors are then introduced as a result of the secondary capacity $C_s$.

Several indirect methods can be used to obtain the effective secondary capacity $C_s$. One way is to determine the frequency $\omega_1/2\pi$ at which parallel resonance occurs across the secondary terminals when the primary terminals are short-circuited. This frequency is that at which $C_s$ and $L_s$ are in parallel resonance, so that

$$C_s = \frac{1}{\omega_1^2 L_s} \tag{47}$$

Another way of getting at the same thing is to measure the ratio of secondary to primary voltage as a function of frequency at frequencies approaching the resonant frequency. This ratio will increase with frequency and reach a maximum when $L_s$ and $C_s$ are in resonance. At some frequency below resonance,

$C_s$ can be deduced from a knowledge of $L_s$ and the step-up ratio according to the relation

$$\frac{\text{Voltage ratio of transformer}}{\text{Step-up ratio of transformer}} = -\frac{(1/\omega C_s)}{\omega L_s - 1/\omega C_s} \tag{48}$$

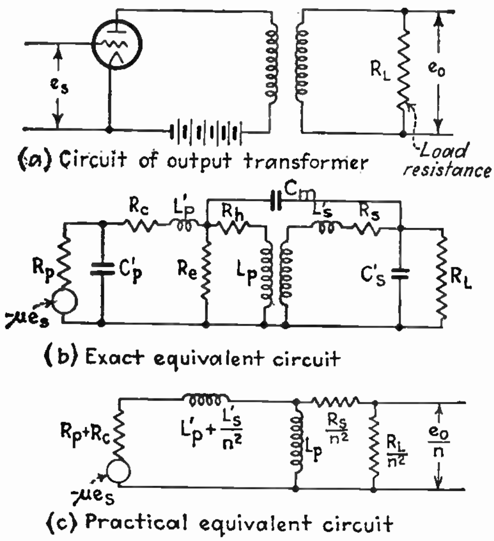
Other indirect methods of determining $C_s$ can be devised when the occasion requires. In measuring $C_s$ it must be kept in mind that the capacity which is effective is the sum of the distributed secondary capacity of the transformer and the input capacity of the tube to which the secondary delivers its voltage. Consequently $C_s$ should be determined with the output tube actually present and operating with its normal load impedance. Any additional measuring equipment, such as vacuum-tube voltmeters, across the secondary must be so arranged as to introduce the minimum possible additional shunting capacity.

With the constants of the transformer determined in the manner described above, it is possible with the aid of the equivalent circuit shown in Fig. 112*c* to predict the curve of amplification as a function of frequency in transformer-coupled amplifiers with an accuracy that is surprisingly high, and which is entirely satisfactory for design purposes.

*Input Transformers or Line-to-grid Transformers.*—These transformers are for the purpose of coupling a microphone or transmission line to the grid of a tube, and differ from interstage transformers only in that the line or microphone resistance replaces the plate resistance of the tube. The primary inductance is proportioned accordingly, and the analysis is the same as in the interstage transformer.

*Output Transformers of Class A Amplifiers.*—Output transformers are employed for coupling the plate circuit of a power amplifier tube to the load impedance that consumes the power. The situation under which these transformers operate is quite different from that in interstage transformers since a step-down ratio is normally called for, and the secondary is shunted by a relatively low-resistance load. Furthermore the power-handling capacity of the transformer is now of importance, whereas in interstage and input transformers the voltages involved are usually extremely small.

The complete equivalent circuit of an output transformer is shown in Fig. 113*b*, but this can for all practical purposes be simplified as shown at Fig. 113*c*. It will be noted that all of the transformer capacities have been dropped because of the fact




NOTATION

$R_p$ = Plate resistance of tube
$R_c$ = D.C. resistance of primary
$L'_p$ = Primary leakage inductance
$R_e$ = Resistance representing eddy current loss
$R_h$ = Resistance representing hysteresis loss
$L_p$ = Incremental primary inductance
$R_s$ = Resistance of secondary winding
$R_L$ = Load resistance
$L'_s$ = Secondary leakage inductance
$C'_p$ = Capacity in shunt with primary (Tube plus transformer capacity
$C'_s$ = Secondary distributed capacity
$C_m$ = Capacity between primary and secondary windings of the transformer
$n$ = Ratio of secondary to primary turns

FIG. 113.—Actual circuit of an output transformer, together with exact and approximate equivalent circuits.

that they are shunted by relatively low resistances and so now have negligible effect. Likewise the eddy-current loss and the hysteresis loss can be ignored, or considered part of the load resistance. The primary and secondary copper resistances are effectively a part of the plate and load resistances, respectively. The significant constants of the equivalent circuit are the primary inductance, which determines the way in which the output drops off at low frequencies, the leakage inductance, which causes the output to drop off at high frequencies, and the ratio of transformation, which determines the equivalent load impedance referred to the primary side of the transformer.[1] These constants can all be measured by the same technique described for interstage transformers, so that the merit of an output transformer and the approximate frequency range which it can be expected to cover under any given set of circumstances can be quickly determined by simple bridge measurements.

In the output transformer it is necessary to consider the maximum alternating voltage which may be applied to the primary without saturating the core and hence producing a highly distorted wave of magnetizing current. Harmonics introduced by the non-linearity of the magnetization curve of the core are always present to some extent. They can be minimized, however, by designing the transformer in such a way that the alternating flux density in the core is small under the conditions of operation, and by making the inductive reactance of the primary high. The extent of the distortion introduced by the non-linear properties of a core iron can be determined by applying a 60-cycle voltage of suitable magnitude and observing the wave shape of the magnetizing current using an oscillograph. In such a test the arrangements should be such that the flux density obtained with 60 cycles is the same as the flux density at the lowest frequency at which it is desired to avoid distortion in the transformer. Thus, if this lowest frequency is 80 cycles and the crest voltage expected is 200 volts, then one must apply $(60/80) \times 200 = 150$ volts crest at 60 cycles to obtain the same flux density. When the transformer is to carry direct current, the appropriate direct-current magnetization must also be provided.

*Push-pull Transformers.*—In Class A amplifiers employing push-pull transformers, the situation is the same as in ordinary

[1] Analysis of the circuit of Fig. 113 shows that the output voltage of a power amplifier drops to 70.7 per cent of its maximum value at a low frequency such that the inductive reactance of the primary is equal to the resistance formed by $(R_p + R_c)$ in parallel with $(R_L + R_s)/n^2$, and also falls to 70.7 per cent of the maximum at a high frequency such that the inductive reactance of $L_p' + L_s'/n^2$ is equal to $R_p + R_c + (R_L + R_s)/n^2$.

Class A single-tube amplifiers except that the effective plate resistance associated with the transformer primary is twice the plate resistance of a single tube, and one takes the ratio of transformation as being from the whole primary to the whole secondary. In push-pull transformers the only direct-current magnetization of the core is that which results from unbalance between the plate currents of the two tubes, and this seldom exceeds 10 per cent of the plate current of an individual tube. The incremental inductance of the primary must be measured under corresponding conditions.

In Class B audio amplifiers the situation is complicated and the circuit details as they affect the transformer have not yet been completely worked out in a systematic manner. In general, the transformer of the driver stage always has a step-down ratio, and behaves very much the same as an output transformer in a Class A amplifier. The output transformer of the Class B stage acts in much the same way as in the Class A case, and the same factors are effective in controlling performance.

# CHAPTER IX

## RECEIVER MEASUREMENT

**50. Determination of Receiver Performance.**—The performance of a radio receiver is measured by using an artificial signal to represent the voltage which is induced in the receiving antenna, and applying this artificial signal to the input terminals of the receiver through a network or "artificial antenna" that simulates the impedance of the actual antenna for which the receiver is to be used (see Fig. 114). The equipment for producing the artificial signal is called a standard signal generator, and consists of a thoroughly shielded oscillator coupled to an attenuating system that is capable of producing known voltages ranging from about 1 microvolt up to perhaps 200,000 microvolts. Provision is made for modulating this voltage to any degree up to 100 per cent and at all modulation frequencies that will be encountered. Unless the nature of the test requires otherwise, however, it is standard practice to modulate the carrier voltage developed by the signal generator to 30 per cent at 400 cycles. A test signal modulated in this way is often referred to as a "standard test voltage."

The output of the radio receiver is determined by measuring the power developed in a non-inductive load resistance that is connected across the output terminals in place of the loudspeaker or other load impedance normally present. The resistance should be that for which maximum undistorted power output is obtained. The output power can be evaluated with the aid of a vacuum-tube voltmeter or a thermocouple instrument, or by using an output meter such as described in Sec. 9 for supplying the load impedance and measuring the power. The normal test output power for broadcast and similar receivers is 50 milliwatts.

The importance of uniform standards for broadcast receivers is such that the test procedure has been prescribed in considerable detail. The standard artificial antenna for making broadcast receiver tests consists of a capacity of 200 $\mu\mu$f, a self-inductance of 20 $\mu$h, and a resistance of 25 ohms, all in series. The output resistance of the signal generator is included as part of this 25-ohm resistance. These values are selected as approximating the constants of the average broadcast receiving antenna. It is also specified that tests should be made at certain carrier frequencies, which are 600, 1000, and 1400 kc if three test frequencies are to be employed, with the addition of 800 and 1200 kc when five test frequencies are used.

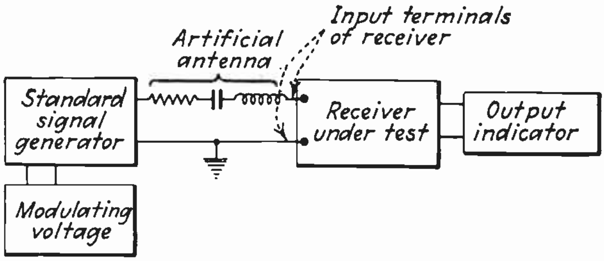



FIG. 114.—Schematic arrangement of equipment for making measurements of receiver performance.

A summary of the more important procedures involved in making receiver tests is given below. These are applied specifically to the case of broadcast receivers but apply with only minor and obvious modifications to all receivers of modulated waves. A complete statement of the various standardized features of receiver tests is to be found in the reports of the Standardization Committee of the Institute of Radio Engineers.[1]

*Selectivity, Sensitivity, Fidelity, and Overload Level of Broadcast Receivers.*—The most important features of broadcast receiver performance are the selectivity, sensitivity, fidelity, and overload level. The sensitivity represents the ability of the receiver to respond to weak radio waves and is defined quantitatively as the amplitude of standard test signal (modulated 30 per cent at 400 cycles) which is required to develop the standard 50-milliwatt receiver output with the volume control set for maximum sensitivity. The measuring procedure hence consists in setting the carrier of the signal-generator voltage to the desired frequency, applying the signal-generator output to the broadcast receiver

[1] See Proposed Standard Tests of Broadcast Radio Receivers, *Proc. I.R.E.*, vol. 18, p. 1282, August, 1930.

through the artificial antenna, tuning the receiver to maximum response, and then adjusting the amplitude of the signal-generator output until the standard 50-milliwatt output is obtained. A typical curve of sensitivity as a function of carrier frequency is shown in Fig. 115.

Fidelity represents the extent to which the receiver reproduces the different modulation frequencies without frequency distortion. Fidelity is measured by setting the carrier of the standard test signal to the desired frequency, tuning the receiver for maximum response, and setting the test voltage so the standard 50-milliwatt output is obtained. The modulating frequency is then varied through the desired range while the degree of modulation is kept constant at 30 per cent, and the variation of output is observed. A typical fidelity curve is shown in Fig. 115.

Selectivity is defined as the degree to which the receiver is capable of discriminating against interfering signals having a carrier frequency differing from that of the frequency to which the receiver is tuned. Selectivity cannot be defined in a single term but must be given in the form of curves such as those of Fig. 115, which show the amplitude of a standard test voltage required to give the standard 50-milliwatt output plotted as a function of the difference between the actual test frequency and the frequency to which the receiver is tuned.

The measuring procedure for determining the selectivity of receivers having full manual volume control consists in first determining the sensitivity when the receiver is tuned to the test frequency, following exactly the procedure used in the sensitivity test. The carrier frequency developed by the signal generator is then varied by known amounts, usually in 10-kc steps, and the amplitude of the standard test voltage is increased as necessary to maintain the standard 50-milliwatt output. The selectivity curve is then plotted as shown in Fig. 115.

When the receiver has an automatic control, the situation is complicated by the fact that the receiver sensitivity depends upon the total radio-frequency voltage applied to the automatic volume control electrode. If the selectivity curve is taken as outlined above in the absence of a desired signal, the results will tend to be for constant receiver gain irrespective of frequency. However, one is interested in selectivity primarily as an indication of the interference to be expected when the desired signal

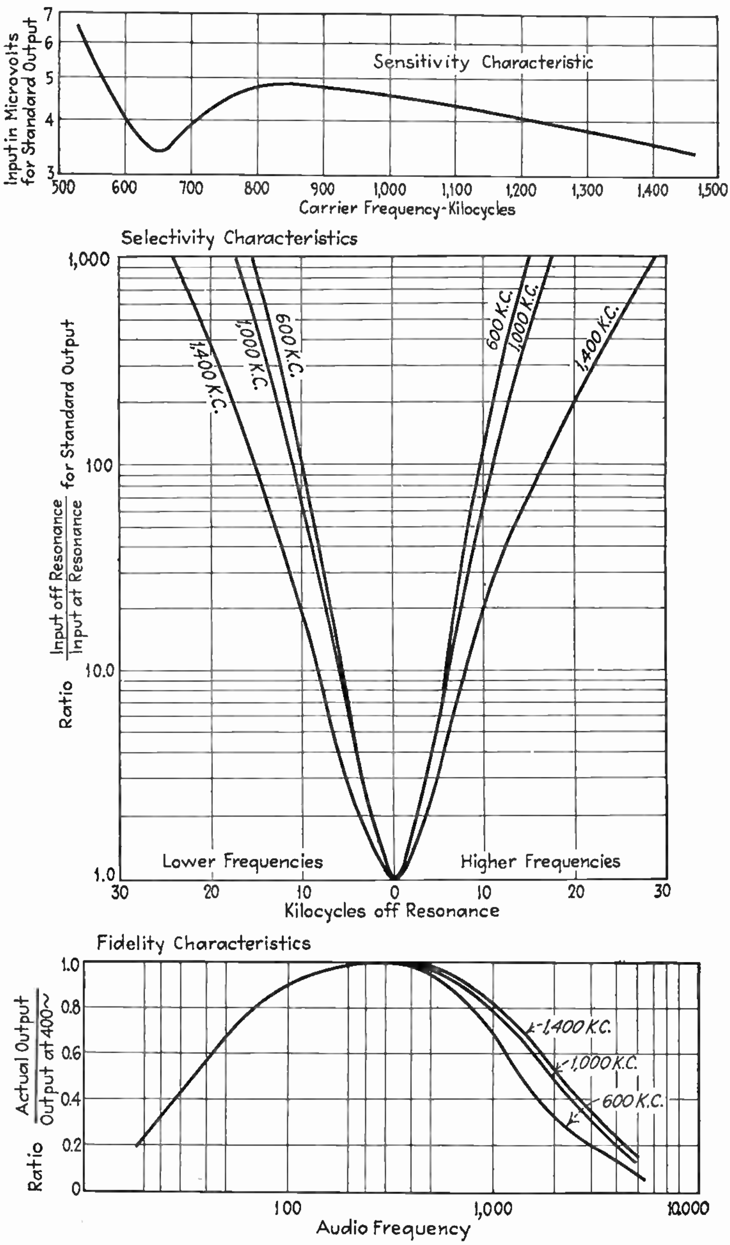



FIG. 115.—Sensitivity, selectivity, and fidelity characteristics of a broadcast receiver having manual volume control.

is being received. The presence of this desired signal alters the radio-frequency voltage acting on the volume control, and so affects the receiver sensitivity. This situation can be simulated by two signal generators, one to supply the "desired" signal, to which the receiver is tuned, and the other to represent the "interfering" signal. The exact details of the testing procedure may be varied in a number of respects, and no standard procedure has been agreed upon. A typical method would be to set the amplitude of the desired signal generator at a value corresponding to the condition for which the test is to be made, after which the modulation of this signal generator is removed so that only the unmodulated carrier frequency of appropriate amplitude is left. The second signal generator is now turned on and arranged as described below to superimpose on the receiver input an additional voltage which represents the interfering signal. This interfering signal generator is given the standard modulation of 30 per cent at 400 cycles. The selectivity is then given by plotting a curve showing the amplitude of the interfering signal required to produce the standard output, as a function of the difference between the frequency of the interfering signal and the frequency of the desired signal.

The two-signal method of measuring selectivity is sometimes used even in receivers having manual volume control in order to take into account the fact that a linear detector produces an apparent increase in the selectivity by enabling the carrier voltage of the desired signal to suppress more or less completely any modulation of a weaker interfering signal.[1] This effect can be observed by noting the reduction in the output produced by the interfering signal as an unmodulated desired signal is turned on and off.

The overload level of a receiver is arbitrarily defined as the output power at which the total r.m.s. value of the harmonic content of the output reaches 10 per cent when the signal generator is sinusoidally modulated to 30 per cent at 400 cycles. The overload level of a receiver is accordingly measured by increasing the output of the standard test signal until analysis of the output wave by one of the methods described in Sec. 38 indicates the limiting harmonic content has been reached.

[1] See "Radio Engineering," p. 319, for further discussion.

**51. Miscellaneous Receiver Characteristics.** *Volume-control Characteristics.*—The effectiveness with which the automatic volume control is able to hold the output constant for signals of different amplitude, and also the delay characteristics of the automatic volume-control arrangement, can be expressed in a curve such as shown in Fig. 116 giving the receiver output as a function of signal amplitude. The procedure for making such a test is to measure the audio-frequency output developed by a receiver as the standard test voltage is varied in amplitude.

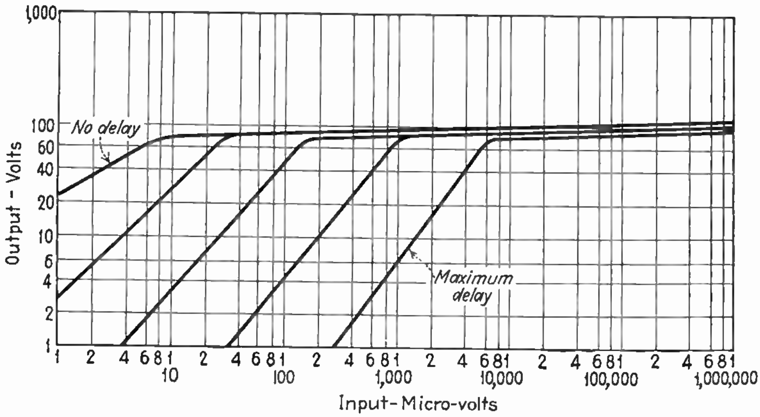



FIG. 116.—Audio-frequency output voltage of a receiver as a function of input signal in receiver having automatic volume control with various "delay" characteristics.

The volume-control setting in addition to controlling the sensitivity may also affect to some extent the selectivity, fidelity, cross-talk, noise, etc. In order to obtain a complete picture of receiver performance it is hence necessary to determine the characteristics with different volume-control conditions, irrespective of whether the volume control is automatic or manual.

*Cross-talk.*—Cross-talk (or cross-modulation) can be measured quantitatively by the use of two signal generators. In making such measurements it is necessary to distinguish between different kinds of cross-talk. The most troublesome type is heard under the following circumstances: The receiver is tuned to a powerful local station—the desired signal—which is so strong as to require a low setting on the volume control. At the same time there is another powerful local station—the "unwanted" signal—operating on a frequency not greatly different from that of the desired station. During the interval in which the desired station is sending out an unmodulated carrier wave, the modulation of the undesired signal will be heard. But if the desired station ceases to radiate its carrier wave, the interfering signal from the unwanted station will disappear. Such cross-talk is caused by the unwanted signal modulating the carrier wave of the desired signal and is the result of third-order modulation products.[1]

The magnitude of this type of cross-talk can be conveniently measured in terms of the percentage of modulation which the unwanted signal produces on the carrier wave of the wanted signal, divided by the percentage of modulation of the unwanted signal. This is known as the cross-talk factor and so is given by the equation

$$\text{Cross-talk factor} = \frac{\text{modulation produced on desired carrier}}{\text{modulation of unwanted carrier}} \tag{49}$$

The cross-talk factor hence represents the effectiveness with which the modulation on the undesired carrier is transferred to the desired carrier wave. The cross-talk factor to a first approximation is independent of the amplitude of the desired carrier and of the percentage modulation of the undesired carrier, and is proportional to the square of the amplitude of the unwanted carrier. The measuring procedure for determining the cross-talk factor consists in setting the desired signal generator at the appropriate frequency and amplitude, and tuning the receiver for maximum response. The modulation is then removed, after which the second signal generator producing the unwanted signal is started up, adjusted to the appropriate frequency and amplitude, and the receiver output noted. The modulation on the unwanted signal is then removed, and the modulation of the desired signal increased from zero up to the point where the same output is obtained as when the unwanted signal is modulated. The cross-talk factor is then calculated from the two modulations with the use of Eq. (49).

[1] The variable-mu tube eliminated most cross-talk problems because of the low third-order modulation it produces at low values of mutual conductance. See Stuart Ballantine and H. A. Snow, Reduction of Distortion and Cross-talk in Radio Receivers by Means of Variable-mu Tetrodes, *Proc. I.R.E.*, vol. 18, p. 2102, December, 1930.

The second kind of cross-talk is produced by heterodyne detection of two signals having a frequency difference lying within the tuning range of the receiver. For example, when one broadcast station is operating on 1400 kc and another on 600 kc, heterodyne detection of the two carrier waves will result in the production of an 800-kc difference frequency which will be heard when the receiver is tuned to 800 kc. The production of such cross-talk requires that the two interfering signals reach the grid of the first radio-frequency tube, and that the characteristic curve of this tube be curved at the operating point. Such cross-talk is normally proportional to the product of the two carrier amplitudes involved, and also to the degree of modulation. It is most readily evaluated quantitatively in terms of the amplitude of a standard test signal of the difference frequency that is required to produce the same output. This form of cross-talk is hence measured by applying to the receiver the signals from two signal generators set to represent the signals involved in the cross-talk. One of these artificial signals is unmodulated, while the other is given the standard modulation of 30 per cent at 400 cycles. The receiver is tuned to the difference frequency and the output observed, after which the unmodulated signal generator is turned off. The carrier frequency of the other signal generator is then reset to the difference frequency, and the amplitude varied until the output is the same as that produced by the cross-talk.

Cross-talk is controlled largely by the point on the characteristic curve at which the radio-frequency tubes of the receiver operate, and so is affected greatly by the setting of the volume control. Cross-talk normally occurs only with strong signals, and when the volume is set to give a low gain, *i.e.*, set to cause the tubes to operate close to cut-off.

*Whistles.*—The production of whistles in a superheterodyne radio receiver can be investigated with the aid of two signal generators. One is normally used to represent the signal to which the receiver is tuned, while the other produces the interfering signal and is varied over a wide range of frequencies in order to determine whether there is any possibility of whistles.

*Hum.*—The hum appearing in a receiver output may be evaluated quantitatively in terms of the hum power which is developed in the receiver output either in the absence of all signals or in the presence of an unmodulated carrier. In interpreting such measurements it must be remembered, however, that the ear is more sensitive to the higher frequency components of the ordinary hum, and the loud-speaker is likewise usually more sensitive to the same components. In order to take this into account it is hence desirable to place between the receiver output and the measuring equipment a network that transmits the various frequency components of the hum in proportion to their importance. No standard network has been agreed upon for this purpose so that it is necessary to work out the details to fit the situation at hand.[1]

*Noise.*—The hissing noise which remains in the output of a receiver when the antenna is disconnected consists of random irregularities arising from thermal agitation, shot effect, etc., and determines the weakest signal which the radio receiver can utilize. The noise energy is uniformly distributed throughout the frequency spectrum and can be evaluated by measuring the noise power with a thermocouple instrument, which indicates the total power irrespective of frequency distribution and also possesses an inertia that averages the power over a brief interval of time. A band-pass filter should be placed in the receiver output to limit the frequency range to a known band. The measurements are often made in the presence of an unmodulated carrier wave of known amplitude since the presence of the relatively large carrier increases the effectiveness with which the relatively small noise voltages are rectified. If desired, the noise level can be evaluated quantitatively in terms of the degree of modulation which this carrier must have to add to the receiver output an amount of power equal to the noise power.[2]

[1] Networks which are often used for making this weighting are described in the following references: Proposed Standard Tests of Broadcast Receivers, *Proc. I.R.E.*, vol. 18, p. 1300, August, 1930; Harold S. Osborne, Review of Work of Subcommittee on Wave Shape Standards of the Standards Committee, *Jour. A.I.E.E.*, vol. 38, p. 1, January, 1919; see p. 6 of this article.

[2] The details for carrying out such measurements are described by Stuart Ballantine, Fluctuation Noise in Radio Receivers, *Proc. I.R.E.*, vol. 18, p. 1377, August, 1930. It is to be noted in this connection that a square-law indicating device (such as a suitably designed thermocouple or vacuum-tube voltmeter) gives a deflection proportional to power so that to add a signal giving the same power output as the noise it is merely necessary to add sufficient signal to double the deflection given by noise alone.

The maximum possible signal-to-noise ratio which can be obtained occurs when all of the noise is a result of thermal agitation in the input circuit of the first tube of the receiver, and the remainder of the receiver contributes no noise. The extent to which a receiver realizes this ideal condition can be determined by disconnecting the antenna and noting the variation in noise output as the input circuit to the first tube is tuned through resonance while the remainder of the receiver is left untouched. The ratio of noise output at resonance to noise output when the input circuit is detuned from the rest of the receiver is a measure of the closeness at which the actual condition approaches the ideal.[1]

*Code Receivers.*—No special standardization technique has been developed for testing code receivers. The same general method as used in broadcast receivers may be followed, with obvious modifications, such as the use of an unmodulated carrier to represent the signal, and the use of a standard output of the order of 1 milliwatt instead of the 50-milliwatt standard for equipment intended to operate loud-speakers.

*Signal Generators in Parallel.*—The problem of superimposing the outputs of two signal generators, as required in cross-talk and some selectivity measurements, presents a difficulty because one terminal of a single generator is always at ground potential. When the signal generators have an output impedance that is constant irrespective of the attenuator setting, the two generators can be connected in parallel. The equivalent output impedance of the combination is then equal to the output impedances of the two generators taken in parallel, while the output voltage of an individual signal generator is less than the output voltage for normal operation by the factor $R_2/(R_1 + R_2)$, where $R_1$ is the output resistance of the signal generator in question and $R_2$ is the output resistance of the other signal generator that is connected in parallel. An alternative arrangement which must be used when the output impedance of the signal generator is variable, and which is also often more convenient in other cases, is illustrated in Fig. 117 and involves the use of separate artificial antennas for each signal generator, with the outputs of these two antennas connected in parallel. Each individual artificial

[1] See F. B. Llewellyn, A Rapid Method of Estimating the Signal-to-noise Ratio of a High Gain Receiver, *Proc. I.R.E.*, vol. 19, p. 416, March, 1931.

antenna is arranged to have twice the impedance of the antenna with which the receiver is supposed to operate; *i.e.*, the resistances and inductances are twice as large as the proper antenna and the capacity is half as great. Since the two antennas are in parallel as far as the receiver input is concerned, the result is equivalent to the desired artificial antenna. It can also be readily shown that the arrangement reduces all output voltages of the signal generators by a factor of 0.5.

**52. Performance of Individual Portions of a Receiver.** *Audio-frequency Amplification.*—The characteristics of the audio-frequency amplifier used in a radio receiver are just the same

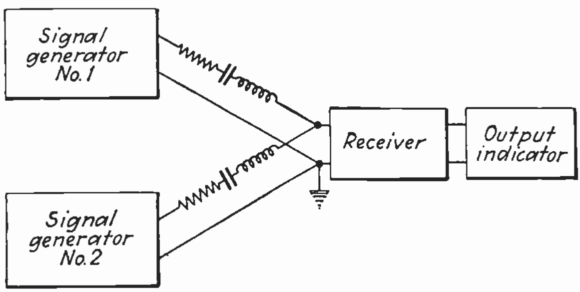



FIG. 117.—Method of superimposing the output voltages of two signal generators upon the input of a radio receiver. The individual artificial antennas have twice the impedance of the antenna with which the receiver is intended to operate.

as though the amplifier was not associated with a receiver. The only special feature that must be considered is the determination of the point at which the audio-frequency amplifying system really begins, and this depends upon the type of detector employed in the receiver. In the case of diode detectors the input voltage to the audio-frequency system is the voltage developed across the load resistance of the diode. In plate detection the audio-frequency voltage is developed in the plate circuit of the tube, so that in making measurements upon the complete audio-frequency amplifier one must either introduce the input voltage in series with the plate circuit of the detector tube or apply to the grid of the tube a voltage $1/\mu$ times as great. It is necessary that an unmodulated carrier voltage of the appropriate amplitude be applied to the grid of the detector tube when making the measurements since the effective plate resistance of the detector tube from the point of view of the audio-frequency voltage is controlled largely by the carrier amplitude of the signal. In the case of grid-leak detection the audio-frequency voltage produced by the detector action is developed across the grid leak and grid condenser and is then applied to the grid of the detector tube. The audio-frequency system therefore starts with the grid of the detector tube, and tests are made accordingly, care being taken to give the grid a negative bias equaling the bias which the normal signal will develop across the grid leak-condenser combination.

*Radio-frequency Amplification.*—The over-all radio-frequency amplification of a receiver may be readily determined by comparing the amplitude of standard test voltage, which must be applied to the receiver input through the artificial antenna to give some convenient output, with the test-signal amplitude that must be applied to the input of the detector tube to give the same output. The gain obtained in this way includes the conversion gain of the first detector of a superheterodyne receiver, and also all regenerative effects that are present.

The gain of an individual amplifier stage from the grid of the input tube to the grid of the following tube is obtainable by the following procedure: A standard test voltage is applied to the grid of the input tube of the stage under investigation (point *aa* of the case shown in Fig. 118) and the carrier amplitude adjusted until the receiver output has some convenient value. The test voltage is then applied directly to the grid of the following tube (point *bb* in Fig. 118), and the amplitude readjusted until the same receiver output as before is obtained. The ratio of the two test voltages is then obviously the gain of the stage. The resonance step-up of voltage between the antenna and the grid of the first tube can be determined in like manner, by first applying the standard test voltage to the receiver through the artificial antenna and adjusting for convenient receiver output, after which the test voltage is applied directly to the grid of the first tube of the receiver and readjusted to give the same output. The ratio of the two voltages is obviously the antenna step-up.

Measurements of amplification of the individual stages and also of antenna step-up, made according to the foregoing procedure, give results which are very near the actual gain existing in the receiver, including regenerative effects. The regeneration is properly allowed for because the principal effect of regeneration in a particular amplifier stage occurs as a result of interchange of energy with a stage of higher power level, and the measuring system outlined does not affect this action since the higher power level stages are allowed to function in a normal manner without any disturbance from the measuring equipment. The energy which a stage of amplification interchanges with another stage of lower power level has negligible effect on the stage of high power level and so need not be allowed for. The impedance changes caused by the measuring equipment introduces no

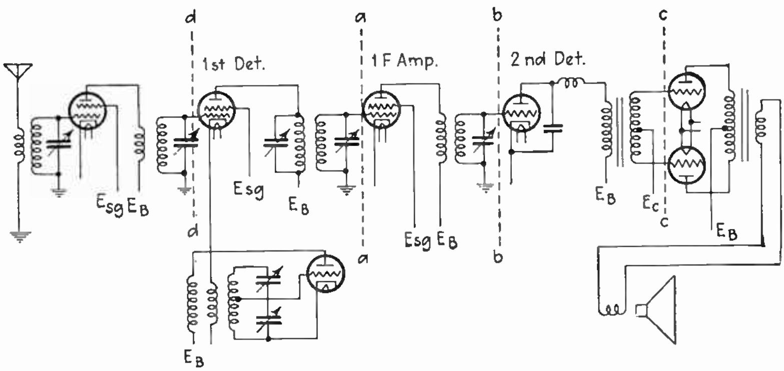



FIG. 118.—Typical superheterodyne receiver, showing points at which test voltages are introduced into receiver to determine performance of component parts.

important error since the measuring equipment is inserted into the receiver in such a way as to affect only the stages preceding the stage under test. The result is that the method gives the gain of the individual stages rather accurately, even though it is too much to expect that the gain of the individual parts of the amplifier obtained in this way will combine to give exactly the observed over-all amplification.

The distortion introduced by the radio-frequency amplifier can be readily determined by making the usual fidelity and overload-level tests with the test voltage applied first at the input and then at the output of the section being investigated, just as above, and then crediting the difference in distortion between the two measurements to the part of the amplifier under test.

*Detector Performance.*—The important characteristics of a detector are the over-all gain of the detector stage, the efficiency of detection as compared with the efficiency obtainable from a perfect rectifier, and the amplitude and frequency distortion which the detector introduces. In making measurements of detection it is necessary to distinguish between frequency changing detectors and the final output detector.

In the final detector the amplification obtained from the input of the detector tube to the input of the first audio-frequency tube is measured by sinusoidally modulating a test signal of known carrier amplitude to a known degree of modulation, applying this to the detector input (point *bb* in Fig. 118), and measuring by the usual audio-frequency amplifier technique the audio voltage which is developed at the grid of the first audio tube[1] (point *cc* in Fig. 118). The conversion gain is then

$$\text{Gain in detector} = \frac{\text{audio output voltage}}{mE_0} \tag{50}$$

where $E_0$ is the amplitude of the carrier voltage applied to the detector input, and $m$ is the degree of modulation of this voltage. The frequency distortion occurring in the detector stage can be investigated by varying the modulation frequency of the test voltage and observing the variation in detector output. The amplitude distortion is likewise determined by investigating the wave shape of the detector output voltage when sinusoidal modulation is used.

The efficiency of detection can be defined as the ratio of the audio-frequency output voltage actually obtained to the audio-frequency output voltage which would be obtained with a perfect rectifier. In a diode rectifier, the maximum audio-frequency output voltage under ideal conditions is a voltage of $mE_0$ developed across the load resistance, so that the efficiency is the audio output voltage divided by $mE_0$, which is also the gain of the detector.

In grid-leak power detectors a perfect rectifier causes a voltage $mE_0$ to be developed across the grid leak. The measuring procedure for determining the detector efficiency is hence to apply a carrier voltage $E_0$ modulated to a degree $m$ to the grid

[1] The characteristics of the detector stage obtained in this way include the characteristics of the audio-frequency coupling device between the detector output and the grid of the first audio-frequency tube.

of the detector, and measure the audio-frequency output voltage developed at the grid of the succeeding tube. The detector grid is then biased with a battery to the same negative voltage that the rectification action develops, and an audio-frequency voltage of the modulation frequency is applied to the grid of the detector and adjusted to an amplitude $E_a$ such that the output at the grid of the next tube is the same as that produced during detection. One then has

$$\text{Efficiency of detection} = \frac{E_a}{mE_0} \tag{51}$$

In a similar manner an ideal plate detector gives an output voltage equal to $mE_0A$ where $m$ and $E_0$ have the same meaning as before, and $A$ is the audio-frequency gain obtained from the detector tube when an audio-frequency voltage is applied to the grid of this tube superimposed upon an unmodulated carrier of amplitude $E_0$, and the output is measured at the grid of the succeeding tube. The efficiency of detection is hence the ratio of the audio-frequency gain of the detector stage to the gain of this ideal stage.[1]

The characteristics of the frequency-changing detector of a superheterodyne receiver are determined by the same general principles. Thus the total conversion gain of the frequency-changing stage is obtained by first applying a voltage of signal frequency obtained from a signal generator directly to the input of the frequency-changing tube (with oscillator in operation) and adjusting the test voltage until convenient indication is obtained at the receiver output terminals. The carrier frequency of the test voltage is now changed to the intermediate frequency and the test voltage then applied directly to the grid of the first intermediate frequency tube and adjusted in amplitude

[1] These procedures for measuring detector performance represent the practical methods that give the information normally desired. Those who have gone into the details of detector theory will note that the "efficiency of detection" as applied in the text to grid-leak and diode detectors lumps the coupling efficiency of the load impedance formed by grid leak and condenser with the true "efficiency of rectification," whereas this is not done in the measuring procedure outlined for plate detection. It is not difficult to measure separately the coupling efficiency and true rectification efficiency of diode and grid-leak detectors, but in ordinary circumstances there is no point in doing so since the coupling efficiency is uniformly high, usually 90 to 98 per cent, and so causes only a relatively small loss.

until the same receiver output is obtained. The two measuring points are hence *dd* and *aa*, respectively, in Fig. 118. The conversion gain is then obviously the ratio of the two test voltages required, provided the degree of modulation is kept constant.[1] Similarly the detector efficiency with plate detection is the ratio of the actual conversion gain to the gain that is obtained by considering the detector tube to be an intermediate frequency amplifier tube. This latter gain is measured *while the local oscillator is operating in normal manner*, by using the technique already given for obtaining the performance of individual amplifier stages.

**53. Signal Generators.**—A signal generator is a device for producing an artificial signal that can be varied in amplitude from about 1 microvolt up to about 200,000 microvolts. It consists of a modulated oscillator which is completely shielded and is provided with an adjustable attenuator for producing known output voltages. The shielding must be practically perfect since it is necessary that the leakage be negligible compared with attenuator outputs as small as 1 microvolt. The attenuator likewise must be carefully designed since it is called upon to produce very small voltages by reducing larger measurable potentials by amounts that can be calculated.

*Shielding of Signal Generators.*—Shielding is accomplished by using copper or aluminum boxes to confine magnetic and electrostatic flux, and by the use of filters to prevent leakage along the wires which must pass through the shield. The problems encountered in shielding an oscillator can be understood by reference to Fig. 119. A simple shield box, with filters in each outgoing lead as shown in Fig. 119*a*, will eliminate most of the field that would otherwise be produced by an unshielded oscillator, but it is always found that there is residual field external to the oscillator which will be present irrespective of the thickness of the shielding box or the number of filter sections

[1] This method will give erroneous results if the method of coupling the local oscillator into the first detector tube causes appreciable local oscillator voltage to be developed across the tuned input circuit to the detector. This is because such voltages affect the conversion gain, and are short-circuited out when the signal generator is connected directly to the grid. See W. A. Harris, "The Application of Superheterodyne Frequency Conversion Systems to Multirange Receivers," *Proc. I.R.E.*, vol. 23, p. 279, April, 1935.

placed in the outgoing leads. The reason is that the shield serves as ground return for the various circuits, and also shields the magnetic flux produced by the coil. The result is that the shield carries an appreciable current which produces potential differences between parts of the shield that give rise to external electrostatic flux and magnetic flux. These potential differences

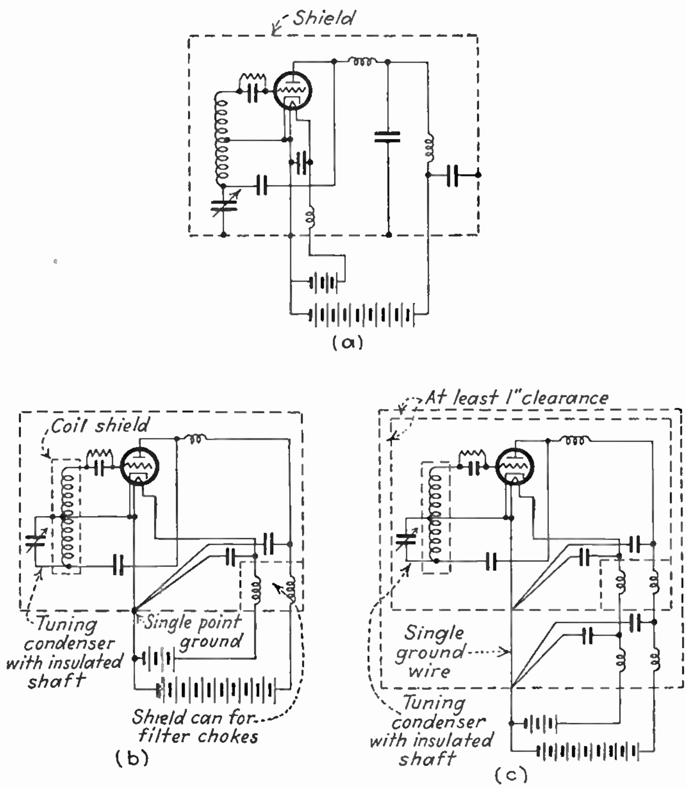



FIG. 119.—Successive steps in obtaining complete shielding of an oscillator.

between parts of the shield also apply voltages between the outgoing leads that limit the maximum filtering effect obtainable, while at the same time stray fields within the shield will induce voltages in the filter chokes.

The remedy for these deficiencies of Fig. 119*a* is shown in Fig. 119*b* and consists in rearranging the grounds so that there is only one grounding point, with *individual* ground wires for each filter condenser, in placing the coil in a separate shield grounded to the main shield only at the common grounding point, and in placing the filter chokes in separate compartments. It is also helpful to make the coil and its shield as compact as possible. When these steps are taken, the only currents carried by the shield are the incidental currents caused by stray couplings and these are so small that the shield is very nearly an equipotential surface. The effectiveness of the filtering in the output leads is greatly increased by the rearrangement of grounds, and also by shielding the filter chokes from the fields inside the shielding box.

The improvements effected in the shielding by using the arrangement of Fig. 119*b* are considerable, but not sufficient to meet the requirements of signal-generator service. The final step is shown in Fig. 119*c* and consists in surrounding the arrangement of Fig. 119*b* with a second shield box which is connected to the inner shield at only a *single point* and is provided with an additional stage of filtering where the outgoing wires leave the outer shield. The outer box should be large enough to clear the inner shield by at least 1 in. It is essential that there be only one ground connection between the two shields, since if there are two grounding points the potential differences between the two grounding points on the inner shield will produce circulating currents in the outer shield that give rise to external fields. It is therefore necessary that the inner shield be mounted upon insulating studs and that the shaft of the tuning condenser be broken with an insulating bushing. The connecting wires in Fig. 119*c* do not destroy the single-point connection because they have a high-impedance radio-frequency choke in the circuit between points that are by-passed to the respective shields.

The details of the filters in the outgoing wires vary according to the circumstances. In the filament circuit the by-pass condensers normally are from 0.1 to 0.5 $\mu$f, and the chokes are designed to have a low direct-current resistance and several hundred microhenrys inductance. It is desirable to use heater-type tubes where possible, since then there is very little high-frequency current to be filtered out of the heater leads. The filter sections in the plate circuit must normally be designed to permit plate modulation of the oscillator from an external oscillator, while at the same time preventing any radio-frequency leakage. This result is accomplished by designing the plate filter as a simple low-pass filter composed of $\pi$ sections having a characteristic impedance that corresponds approximately to the effective load resistance which the plate circuit of the oscillator tube offers. The cut-off frequency is usually made appreciably higher than the highest modulation frequency in order that the transmission may be relatively uniform over the normal range of modulating frequencies in spite of the fact that the filter is not provided with terminal half sections. The danger of leakage through the plate filter can be greatly reduced by using a grounded-plate oscillator circuit as shown in Figs. 120 and 121 since then very little radio-frequency voltage is applied to the plate lead. With heater-type tubes and a grounded-plate oscillator circuit, adequate filtering is usually obtained with two filter sections in each lead, one at each shield.

The shield boxes can be built up from moderately heavy sheet metal. The inner box is usually made of copper with soldered seams and a tight-fitting lid sometimes hinged and sometimes clamped on with thumb screws. The outer case is commonly built up from sheet aluminum mounted on an angle-iron frame. Copper is too soft for use here, and brass has too much resistivity to serve as an effective shield. The lid on the outer box is either clamped on with thumb screws or hinged and arranged to give effective sealing.

*Signal Generators Having Resistance Attenuators.*—The most common method of deriving a known voltage from the shielded oscillator is by the use of a ladder-type attenuator composed of resistance elements and supplied with a known current, as shown in Fig. 120. The output from such an attenuator is varied in large steps by means of the switch, while continuous adjustment between steps is obtained either by varying the input current or by using a calibrated slide-wire potentiometer for the final shunting resistance of the ladder.[1] It is assumed that the attenuator output can be calculated from the resistance elements of the attenuator, which means that great care must be taken in the design and arrangement to avoid errors at high frequencies. The attenuator input is supplied from a pick-up coil coupled to the oscillator coil but insulated from the inner shield so that the

[1] For the design of a non-inductive potentiometer for this purpose see J. R. Bird, The Design of Radio-frequency Signal Generators, *Proc. I.R.E.*, vol. 19, p. 438, March, 1931.

voltage drop produced in the leads between coil and attenuator by the attenuator currents will not affect the relative potentials of the inner and outer shielding cans. The input current to the attenuator is controlled by a rheostat and measured with a thermocouple instrument, with a filter arranged as shown in Fig. 120 placed in the leads between the thermocouple and its galvanometer so that the latter may be mounted on the outer shield. The leads between the pick-up coil and the attenuator are preferably arranged in the form of a twisted pair or a concentric cable to minimize the voltage drop along them. The pick-up and oscillator coils are also preferably not too close together and are arranged so that the ends which face each other are both at ground potential. This reduces electrostatic coupling between the coils to a minimum, with a consequent reduction in the capacity currents that circulate out along the ground wire of the pick-up coil to the outer shield, and then back to the inner shield through the ground connection between shields.

The attenuator itself must be designed to have an output impedance of 5 to 10 ohms in order that it may operate satisfactorily with the usual artificial antenna. Because of this low impedance the principal shielding problems in the attenuator involve inductive rather than capacitive couplings because of the relatively large currents, and low impedance to ground. The resistances must be wound with wire sufficiently small to have negligible skin effect, and must have negligible phase angle. The mica-card, figure-eight, and reversed-loop systems illustrated in Fig. 59 are commonly used. The points at which the shunting elements are connected to the return wire of the attenuator should preferably be distributed along this wire so that the portions of the return wire carrying heavy current, and so having a large voltage drop per unit length, will be separated from the attenuator output when the output voltage is very small, *i.e.*, when the attenuator switch is connected to the left-hand side of the attenuator in Fig. 120.

The most satisfactory method of mounting the attenuator and its accompanying switch is illustrated in Fig. 120. It involves inclosing the attenuator in a heavy copper or aluminum shield which is sectionalized to provide shielding between parts operating at widely different power levels. The switch is mounted in a separate compartment as shown, and arranged

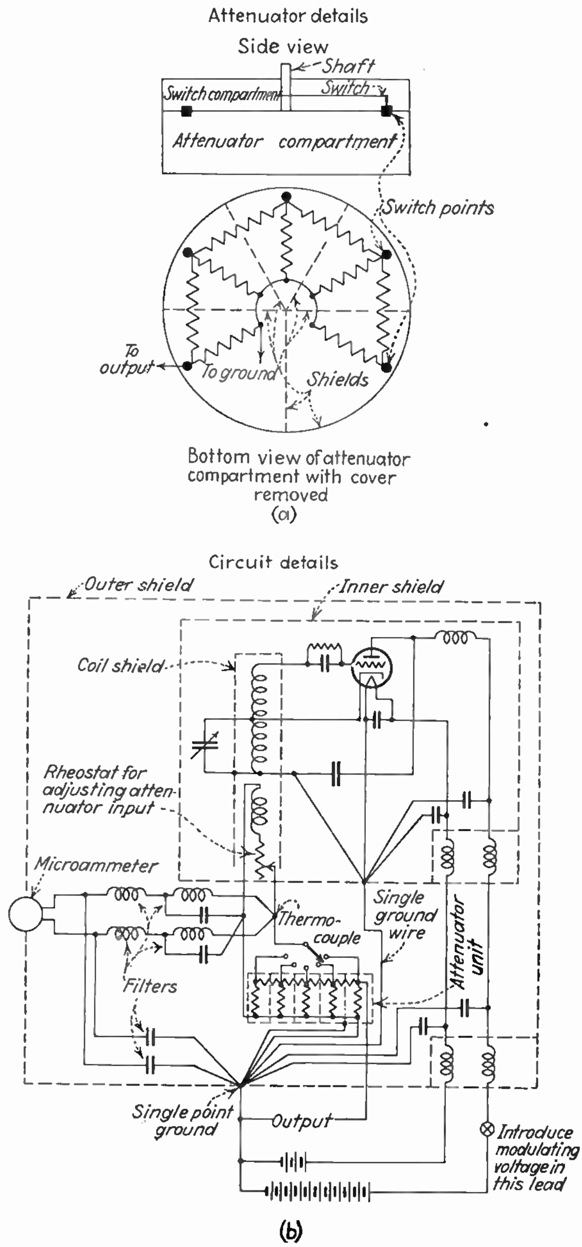



FIG. 120.—Signal generator of the resistance attenuator type.

so that the coupling between the switch arm and the contact points not in use is as low as possible. The shield is grounded at only a single point.

The best resistance attenuators have negligible frequency error up to about 2000 kc, with an error that does not exceed 20 per cent at 25,000 kc when the output is 1 microvolt. With larger outputs the error at high frequencies is correspondingly less. The great advantages of the resistance attenuator are that it is direct reading in output voltage, and that no correction is normally required for frequency.

*Signal Generators Having Mutual-inductance Attenuators.*—In this type of signal generator the output voltage is the potential which is induced in a pick-up coil coupled to the oscillator coil. The ratio of the induced potential to voltage across the oscillator coil is independent of frequency (assuming that the oscillator coil has a reasonable $Q$) and is proportional to the mutual inductance. The attenuation can therefore be controlled by varying the spacing between the two coils, and can be evaluated by either calculating or determining experimentally the mutual inductance as a function of spacing.

A particularly satisfactory form of mutual-inductance attenuator is illustrated in Fig. 121 and involves mounting the oscillator and pick-up coils in a heavy copper tube.[1] The field produced in such a tube by the oscillator coil dies away exponentially with distance, provided the coils are not too close together. The attenuation in decibels is consequently a linear function of the spacing and can be calculated with a high degree of accuracy from the dimensions involved. The signal-generator output is most conveniently indicated in decibels attenuation below some arbitrary output (commonly the position where the output is approximately 1 volt) rather than in microvolts. The actual output at the zero-level position is readily measured with a vacuum-tube voltmeter.

[1] Attenuators of this type may be called piston attenuators. They were developed in the form herein described by H. A. Wheeler of the Hazeltine Corporation. The author is particularly indebted to Mr. Wheeler and the Hazeltine Corporation for making information on these attenuators available for use here. A paper by Harnett and Case describing piston-type attenuators in greater detail, and also describing capacity attenuators using the piston principle, can be expected to appear in the *Proceedings* of the Institute of Radio Engineers toward the end of 1935.

When the oscillator and pick-up coils are close together, the size of the coil has some effect on the rate of attenuation with

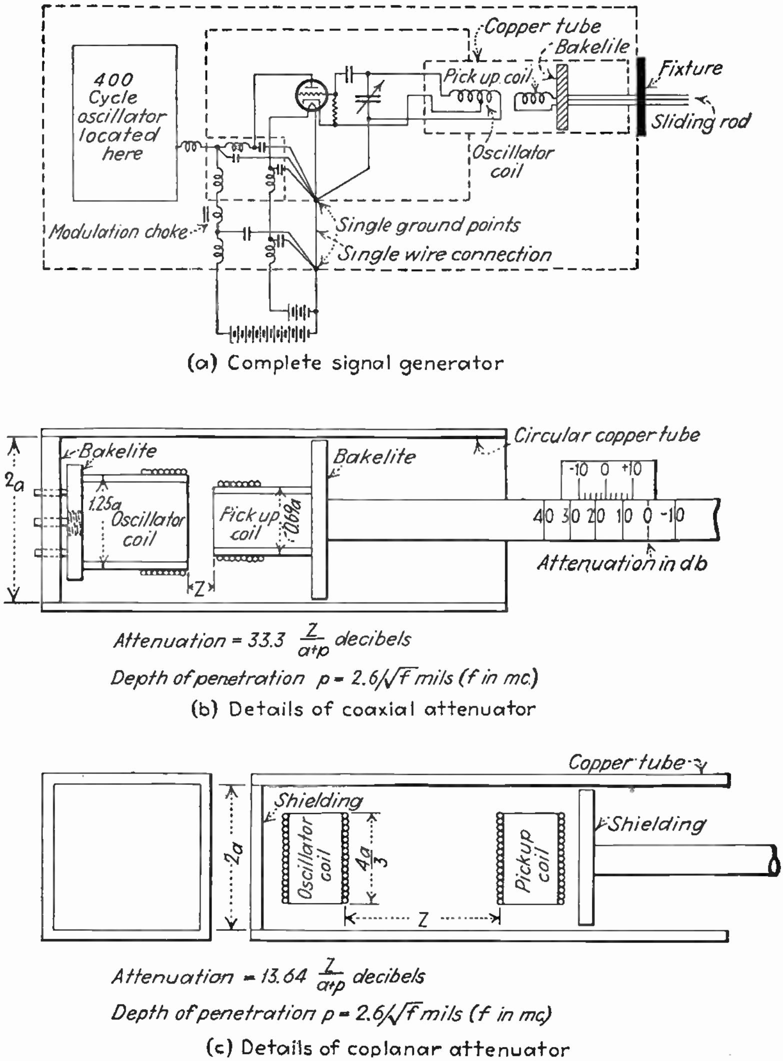



FIG. 121.—Signal generator of the mutual-inductance attenuator type.

spacing. The proportions illustrated in Figs. 121*b* and 121*c* minimize this effect and keep the rate of attenuation substantially constant up to the point where the coils are so close together that the reaction of the pick-up coil on the oscillator is excessive.

The rate at which attenuation increases with spacing depends only upon the "effective" inside dimension of the copper tube, provided the coils are not too close together. In evaluating the effective inside dimensions one includes the "depth of penetration" of the current as part of the inside area. Thus the effective radius of a cylindrical tube is $a + p$, where $a$ is the measured inside radius and $p$ is the depth of penetration, while in a square tube having a side length $2a$ the effective side length is $2a + 2p$. Since the depth of penetration varies as the square root of frequency, the law of attenuation varies slightly with frequency. With practical proportions the effect is small, however. Thus with a tube 2 in. in diameter, increasing the frequency from 1000 to 25,000 kc alters the attenuation by only ¼ db in 100 db.

The most convenient arrangement of oscillator and pick-up coils is the coaxial one illustrated in Figs. 121*a* and 121*b*. The coils can be readily changed by unscrewing the fixture in the outer case, and handling the oscillator coil through the resulting hole with the aid of a bakelite rod threaded to screw into a corresponding hole in the coil-mounting base. The principal source of errors in mutual-inductance attenuators with coaxial coils are capacitive couplings and departures from axial symmetry. Both of these effects produce fields that attenuate less rapidly with spacing than does the desired "axial" coupling and so become more detrimental with increasing attenuation. Capacitive couplings can be minimized by grounding the ends of the coils facing each other, as is done in Fig. 121*a*. The lead wires associated with the coils are the principal cause of departure from axial symmetry, and become of increasing importance the fewer the number of turns in the coil. Their effect can be minimized by keeping the going and return wires close together, and arranging the oscillator lead wires in a plane which contains the axis of symmetry, while arranging the wires to the pick-up coil in a plane also containing the axis of symmetry, but at right angles to the first plane.

Another form of mutual-inductance attenuator, having coplanar coils in a square tube, is shown at Fig. 121*c*. With this arrangement the coupling is caused by departure from axial symmetry. The lead wires hence introduce no trouble but the capacity coupling for small spacings is greater than in Fig. 121*a*. The inductive coupling made use of in Fig. 121*c* decreases more slowly with spacing than any other form of coupling, either capacitive or inductive, obtainable in a closed tube. Therefore incidental couplings of other forms become less detrimental with increasing attenuation in this type of attenuator.

The internal impedance of a mutual-inductance type of attenuator is the impedance represented by the pick-up coil. The inductance of this coil must hence be considered as part of the inductance of the artificial antenna. If the load circuit is resonant, there will be a resonant rise of voltage in the inductance and the terminal voltage will be higher than the attenuator calibration indicates, although the voltage actually acting in series with the circuit is correct.

The simplicity, wide frequency range, and ease with which coils may be changed make mutual-inductance signal generators of the type illustrated in Fig. 121 particularly convenient for general receiver testing. When the attenuation scale is obtained by calculation rather than experiment, the signal generator cannot be expected to have the accuracy of resistance attenuators at broadcast and lower frequencies, and the mutual-inductance attenuator also has the disadvantage that the output corresponding to zero level varies somewhat with frequency. This type of signal generator is, however, entirely satisfactory for making ordinary receiver measurements, and is the best form of signal generator that has been devised for use at extremely high frequencies.

*Second-harmonic Type of Signal Generator.*[1]—In this type of signal generator the known voltage is produced by applying a carrier wave of half the desired frequency to the grid of a full-wave square-law detector. When this is done, the plate current of the detector will contain among other things a small second harmonic of the applied frequency, and also a rectified d-c current. It can be readily shown that the crest amplitude of this second-harmonic current is exactly equal to the rectified d-c current, so that a measurement of the latter by a microam-

[1] Signal generators of this type were developed by W. Van B. Roberts. For a further description see W. F. Diehl, A Standard Microvolter Using the Second Harmonic Principle, *Electronics*, vol. 5, p. 230, July, 1932.

meter gives the amplitude of the second-harmonic current with very high precision.[1] A known small radio-frequency voltage can therefore be produced by placing a low resistance, such as 1 ohm, in the plate circuit of the detector tube. The input voltage is then varied until the second-harmonic current, as measured in terms of the rectified plate current, has the value required to develop the desired voltage when flowing through the 1-ohm resistance in the plate circuit. It will be noted that all shielding and attenuator problems are avoided because the oscillator works at half the frequency of the output, and the amplitude of the output current is measured directly.

A practical circuit arrangement for employing the second-harmonic principle in a signal generator is shown in Fig. 122. The rectified plate current, *i.e.*, the change in current caused by the application of the oscillator voltage to the modulator, is measured with a direct-current microammeter provided with an arrangement for balancing out the normal plate current. In

[1] The theory of the second-harmonic signal generator follows: When a signal voltage $e$ is applied to the grid of a tube having a full-wave square-law characteristic, the plate current that flows is

$$i = a_0 + a_1 e + a_2 e^2$$

where $a_0$, $a_1$, and $a_2$ are tube constants. When the applied voltage is a modulated wave,

$$e = E(1 + m \cos vt) \sin \omega t$$

then the current that flows is

$$\begin{aligned} i &= a_0 + a_1 E(1 + m \cos vt) \sin \omega t + a_2 E^2 (1 + m \cos vt)^2 \sin^2 \omega t \\ &= a_0 + a_1 E(1 + m \cos vt) \sin \omega t + \frac{a_2 E^2}{2}\left[1 + \frac{m^2}{2}\right] \\ &\qquad \left[1 + \frac{2m \cos vt}{1 + (m^2/2)} + \frac{m^2 \cos 2vt}{2[1 + (m^2/2)]}\right]\left[1 - \cos 2\omega t\right] \qquad (52) \end{aligned}$$

The first term represents the steady plate current present when no signal is applied, while the second term is a current that reproduces the applied voltage. The third term contains two components, a modulated second harmonic of the applied carrier, and a direct-current component that represents the rectified current (*i.e.*, change in d-c current). It will be noted that these two components are equal irrespective of the modulation, although both increase as the degree of modulation $m$ of the applied voltage is increased. It will be observed that the second-harmonic current has a degree of modulation that is $4/(2 + m^2)$ of the modulation $m$ of the applied wave, and that in addition there is a second-harmonic distortion component of the modulation that is $m/4$ of the desired modulation.

this balancing arrangement, the resistance $R_1$ must be several hundred times the resistance of the microammeter if the meter is to indicate the rectified plate current without correction for the shunt formed by $R_1$. A balanced modulator arrangement is employed in order to reduce as far as possible the oscillator frequency voltage developed across the output resistance $R$. By using radio-frequency pentode tubes and balancing the input to the two tubes very carefully with the condenser $C_1$, it is possible to reduce the oscillator frequency voltage appearing across the resistance $R$ to a very small value. The oscillator

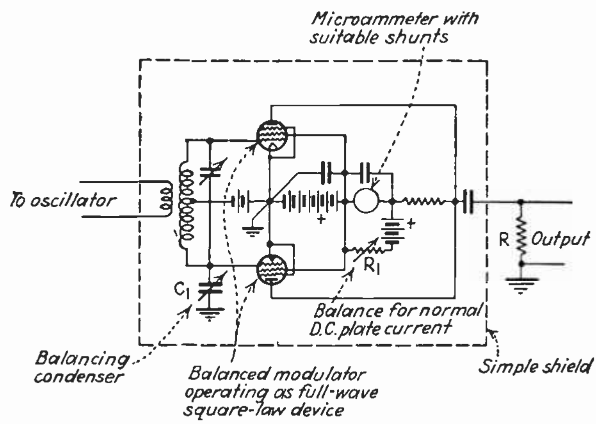



FIG. 122.—Practical form of second-harmonic type of signal generator.

for providing the input to the balanced modulator should be reasonably well shielded, and must be provided with some form of uncalibrated output control for varying the input to the modulator.

The modulation of the second-harmonic output is $4/(2 + m^2)$ times the degree of modulation $m$ of the oscillator. In addition there is a second harmonic of the modulation frequency that is $m/4$ of the desired modulation. The presence of modulation on the input wave does not affect the equality of rectified d-c current and peak amplitude of the second-harmonic carrier, even though it increases the amplitude of both.

The second-harmonic type of signal generator has the very important advantages of requiring practically no shielding, of measuring the output voltage directly rather than depending upon an attenuator, and of being equally satisfactory at all frequencies, including the very highest. The possible accuracy is accordingly greater than obtainable with any other type of signal generator. At the same time there are the disadvantages that the modulation is not distortionless, that along with the desired output voltage there are also voltages of other frequencies which are often much larger than the desired output, and there is finally the necessity of continually checking the balancing adjustment of the microammeter to correct for very small drifts in d-c plate current of the modulator tubes. Thus, although the second-harmonic standard-signal generator is the simplest and most accurate of all signal generators, it is not the most convenient to use.

*Miscellaneous Comments.*—The actual details of signal generators can be varied in many respects. Toroidal coils are sometimes employed, thus making a coil shield unnecessary. Oscillators of the conventional type having amplitude control (see Sec. 71) are used in at least one type of signal generator, and various systems of modulation can be employed. The input to the attenuator is measured with a vacuum-tube voltmeter instead of thermocouple instrument in some signal generators. These and other variations which do not affect the effectiveness of the shielding or the accuracy of the attenuator may be introduced as found convenient.

The oscillator frequency of a signal generator must be stable and as free as possible from frequency modulation. From this point of view it is hence desirable to employ a modulated amplifier arrangement, but this is seldom done because of the complications that result.

Signal generators are normally provided with a self-contained audio-frequency oscillator for producing 30 per cent modulation at 400 cycles. This oscillator is always located in the space between the inner and outer shielding boxes and is arranged so that one may, by means of a switch, obtain the modulation either from this internal oscillator or from an external source of modulating voltage. In the mutual-inductance type of signal generator illustrated in Fig. 121, the modulating oscillator should be located at the opposite end of the inner box from the attenuator in order to minimize troubles from stray couplings. Plate modulation is usually employed in commercial equipment.

Where a wide frequency range is to be covered, plug-in coils are used. Since it is desirable that the tuning-condenser capacity be different for the various frequency ranges, it is often found convenient to use for the tuning condenser a three- or four-gang broadcast condenser in which the different sections are cut down to the appropriate number of plates for the various frequency bands. Each condenser section is connected to an individual jack in the coil base, and the coil forms and their plugs so wired that, upon insertion of the coil, connection is made to the condenser section or sections that should be used with that coil.

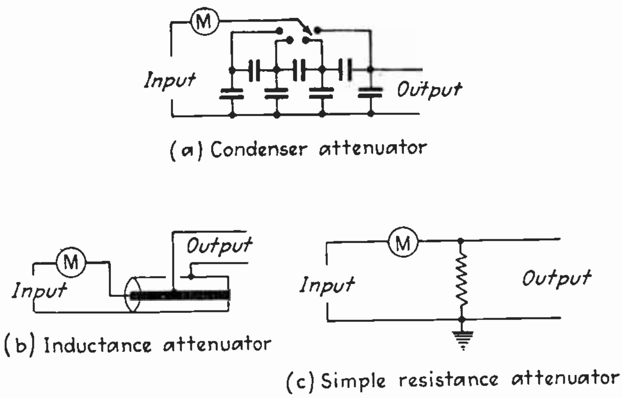



FIG. 123.—Miscellaneous types of attenuators.

Attenuator systems which are occasionally used in signal generators are shown in Fig. 123. The capacity attenuator of Fig. 123*a* is similar to a resistance attenuator except that capacitive reactances now replace the resistance arms. This attenuating system has some advantages, particularly at high frequencies, but suffers in accuracy from the fact that when the capacitive reactance is low the effective series resistance and inductance of the condensers will commonly introduce appreciable errors of unknown magnitude. The inductance attenuator of Fig. 123*b* develops its output voltage as the drop produced by passing a known current through the inductance formed by a short section of a concentric transmission line.[1] The usefulness of this arrangement is limited, however, by the fact that the minimum voltage that can be produced with currents that can be readily measured by a thermocouple is not small enough for

[1] See Hull and Williams, *Phys. Rev.*, vol. 25, p. 147, February, 1925.

many signal-generator applications. Furthermore this type of attenuator does not lend itself readily to covering wide ranges in output voltage. The arrangement of Fig. 123*c*, in which the output voltage is obtained by passing a known current through a known resistance, is capable of developing voltages down to about 50 to 100 microvolts with accuracy. The limit is set by the fact that the minimum current that can be read accurately with a vacuum thermocouple and a portable micro-ammeter is about 500 $\mu$a, while the lowest non-inductive resistance practicable at high frequencies is of the order of 0.1 to 0.2 ohm.

*Accuracy Checks.*—The best method of determining the accuracy of a signal generator is probably by comparing its output voltage with the known output developed from a carefully made second-harmonic type of signal generator. An alternative is to measure the output voltage of the signal generator, using the same technique as that described in Sec. 59, for measuring the voltage induced in an antenna.

An excellent way of detecting errors in an attenuator is to check the lower output ranges against the higher ranges. This can be done with the aid of a sensitive radio receiver in which the final detector has been calibrated to act as a vacuum-tube voltmeter to measure accurately carrier amplitudes at the detector grid over a voltage range of at least 10 to 1. The procedure is illustrated by the following example: The unmodulated signal-generator output voltage is set to some large output, say 10 mv., and coupled to the receiver through an artificial antenna. The receiver gain is adjusted until the carrier at the grid of the detector is as large as the tube can measure, say 10 volts. The signal-generator output is now reduced to 1 mv, and the voltage at the detector again read. The ratio of the two detector readings gives the ratio of signal-generator outputs at the 1- and 10-mv settings and should, of course, be exactly 10. The receiver gain is next increased so that 10 volts appears at the detector with the 1-mv input, after which the signal generator output is set at 0.1 mv and the procedure repeated, and so on. It is possible in this way to check the attenuator step by step, and thereby detect the existence and magnitude of errors that may be present. In making such tests it can normally be assumed that the high output levels are correct since errors when present are always due to spurious couplings and voltage drops which are so small as to be unimportant except at low output levels.

The effectiveness of the signal-generator shielding can be determined by connecting a three- or four-turn coil to the input terminals of a sensitive receiver and exploring for pick-up. The pick-up should be negligible when the coil is at least several inches away from the case, and there should be no pick-up when the coil is brought in the proximity of the leads which go through the generator case. When the signal generator is properly shielded, a sensitive receiver connected to the attenuator posts will show negligible response when the attenuator is set for the smallest possible output voltage and its output posts are short-circuited.

## CHAPTER X

## OSCILLATOR, POWER-AMPLIFIER, AND MODULATION MEASUREMENTS

**54. Voltage and Current Relations in Class C Amplifiers and Power Oscillators.**—Typical oscillograms showing the voltage and current relations that exist in ordinary Class C power amplifiers are shown in Fig. 124. The voltage acting on the grid of the tube consists of a negative-bias voltage $E_C$ greater than the cut-off bias, upon which is superimposed an alternating signal voltage having sufficient amplitude to make the instantaneous grid potential positive during a portion of each cycle. Similarly the voltage actually present at the plate of the tube consists of the plate-supply potential $E_B$ minus the alternating voltage drop $E_{AC}$ that is developed across the tuned load or "tank" circuit. The relations normally existing in the plate circuit are such that the minimum plate potential indicated as $E_{\text{min}}$ in Fig. 124 is only a small fraction of the plate-supply potential $E_B$, and is approximately the same as the maximum positive potential $E_+$ of the grid. The phase relations are such that the minimum plate potential comes at the same instant as the maximum positive grid potential. The alternating voltages at the grid and plate are substantially pure sine waves since they are developed by resonant circuits.

The current that flows at any instant is the result of the joint action of the potentials acting at that instant on the grid and plate electrodes; and if complete characteristic curves of the tube are available, the current oscillogram can be readily derived from the voltage waves. The total space current $I_p + I_g$ is a section of a peaked sine wave lasting somewhat less than half a cycle. The division of this total current between the plate and the grid depends upon the relative potentials of these electrodes, with the plate getting all or nearly all of the current except when the instantaneous grid potential is equal to or greater than the minimum plate potential, when the grid current becomes appreciable.

The most satisfactory procedure for analyzing the conditions existing in a Class C amplifier is to determine the grid- and plate-voltage oscillograms experimentally, and from them derive the current oscillograms with the aid of characteristic curves of the tube. The oscillogram of plate voltage can be derived by determining the plate-supply potential with a direct-current voltmeter and measuring the minimum plate potential $E_{min}$

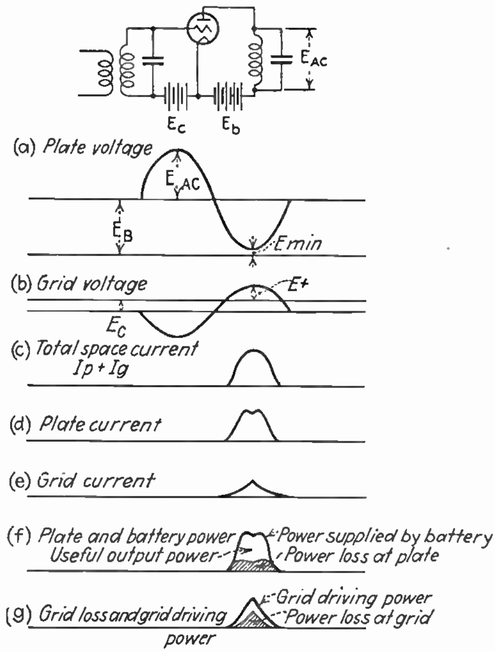



FIG. 124.—Oscillograms showing voltage, current, and power relations in a typical Class C amplifier.

directly with a suitable trough voltmeter of the type described in Sec. 7. The alternating plate potential is then the difference between $E_B$ and $E_{min}$. This method of deriving the oscillogram is always to be preferred to measuring the alternating voltage directly, since the important part of the characteristic is $E_{min}$ and this cannot be determined accurately if it is obtained as the difference of two large and nearly equal quantities. The grid-voltage oscillogram can be derived from a knowledge of the grid-bias voltage, combined with an experimental determination of $E_+$ made with a vacuum-tube voltmeter arranged to measure positive peaks as described in Sec. 7. The exciting voltage is then equal to the sum of the bias potential and $E_+$. This method is preferable to measuring the exciting potential directly since it accurately determines the important quantity $E_+$.

The crest voltmeters used to measure $E_{\text{min}}$ and $E_+$ are illustrated in Fig. 125 and must be arranged to introduce the minimum possible capacity. The trough meter for measuring $E_{\text{min}}$ requires particular consideration since the filament of this tube must be operated at a high potential above ground. This situation can be best handled by lighting the filament from small

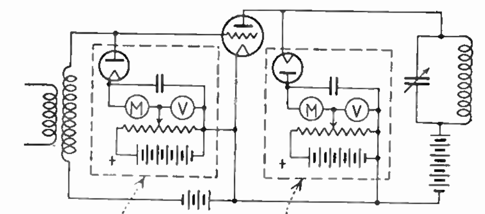



FIG. 125.—Class C amplifier provided with peak voltmeters arranged to measure $E_{\text{min}}$ and $E_+$.

dry batteries which are strapped to the tube base, or by the use of a filament transformer designed to have an especially low capacity between primary and secondary. The tube must likewise be capable of withstanding a high back-voltage, which in the case of the plate trough meter is approximately twice the plate-supply voltage. Of the available tubes, the most satisfactory are the high-voltage half-wave rectifier tubes, Types 878 and 879. In these tubes the plate lead is brought out through the top of the glass bulb so that the electrostatic capacity is low and the voltage rating very high. The allowable space current is very low, but this makes no difference in peak voltmeters that are used to read large voltages.

*Complete Characteristic Curves of Tubes.*—The experimental determination of tube characteristics in the positive grid-potential region is complicated by the fact that the instantaneous plate loss in the tube is great enough to destroy the tube in a few seconds of continuous operation. Damage does not occur in actual Class C operation because the grid is positive only a small fraction of the time, and the average loss is not excessive.

The determination of grid characteristics when the grid is positive therefore involves taking the characteristic "on the run."

One way of doing this is illustrated in Fig. 126[1] and involves first applying the desired plate potential on the tube with zero grid bias. A large condenser $C$ is then charged to a suitable positive potential $E$ and applied to the grid of the tube by means of switch $S$. As the condenser discharges through the resistance $R$ and the grid of the tube, the positive grid potential varies from $E$ to zero so rapidly that the energy dissipated in

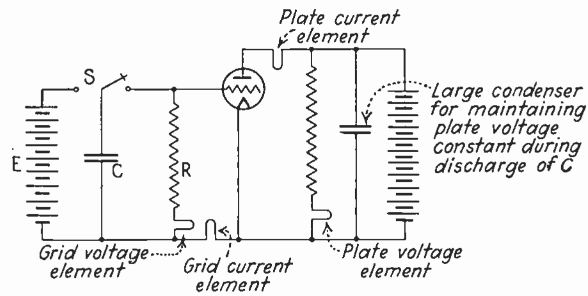



FIG. 126.—Oscillographic method of recording tube characteristics in the positive grid region, using condenser discharge.

the tube during the condenser discharge is not excessive. The variation of plate current, grid current, grid voltage, and plate voltage can be recorded on an oscillogram, and by repeating this procedure a number of times for different plate potentials a complete set of characteristic curves can be obtained.

A second experimental method of obtaining complete tube characteristics is illustrated in Fig. 127. Here the grid is biased well beyond cut-off and a 60-cycle grid exciting voltage is employed, while the plate potential is held as nearly constant as possible by the use of a large plate by-pass condenser. The maximum positive grid potential can be read by a positive crest meter, while minimum plate potential (which is the plate voltage when plate current is maximum) can be read on a trough meter as shown. The plate and grid currents at the crest of cycle are then obtained with the aid of an oscillograph. The tube losses can be kept low by using a very high grid bias,

[1] See H. N. Kozanowski and I. E. Mouromtseff, Vacuum-tube Characteristics in the Positive Grid Region by an Oscillographic Method, *Proc. I.R.E.*, vol. 21, p. 1082, August, 1933.

since then the plate current flows for only a small fraction of each cycle, and the average plate loss is small. This method has the advantage over that illustrated in Fig. 126 in that the action repeats regularly and so can be observed in the viewing screen of the oscillograph. Furthermore one may measure the deflections directly on the viewing screen and avoid all photographic work. If desired, a cathode-ray oscillograph tube employing magnetic deflection may be used to indicate the plate current. When the voltage regulation of the plate battery

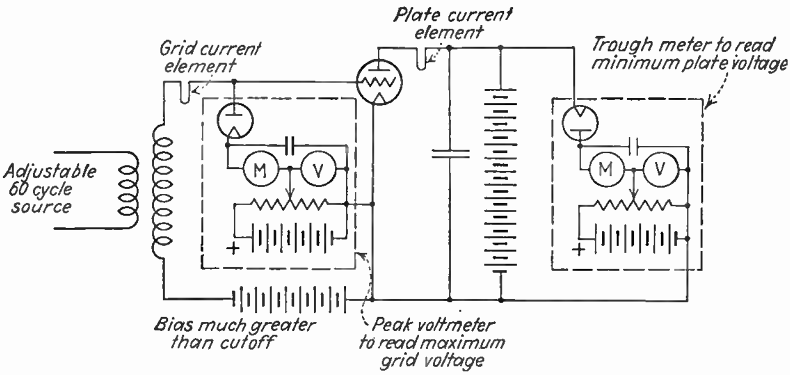



FIG. 127.—Oscillographic method of recording tube characteristics in the positive grid region, using an alternating grid voltage.

(including the effect of its shunting condenser) is extremely good, the trough meter in the plate circuit of Fig. 127 may be omitted.

A cathode-ray tube can also be used to give the characteristic curves of the tube directly as described in Sec. 79.

*Screen-grid Tubes and Power Oscillators.*—The voltage and current relations existing in screen-grid Class C power amplifiers are the same as with triode tubes except for minor modifications introduced by the presence of the screen grid. In particular the operating conditions are normally such that the maximum positive control grid potential $E$ is approximately equal to the screen potential. The minimum plate potential $E_{min}$ must also not be less than the screen potential so that the screen grid will not attract secondary electrons from the plate.

Power oscillators have exactly the same voltage and current relations as Class C amplifiers, since the power oscillator is fundamentally a Class C amplifier in which the exciting voltage for the tube is derived from the output power. It is hence not necessary to consider oscillators separately since the same analysis and measuring methods that have already been described apply without change.

**55. Power Relations in Class C Amplifiers and Power Oscillators.**—The power relations existing in Class C amplifiers can be derived from voltage and current oscillograms such as illustrated in Fig. 124. The power supplied by the direct-current plate potential at any instant is equal to the product of supply potential $E_b$ and instantaneous plate current, and so varies as shown at Fig. 124*f*. The average value over the cycle represents the power which the direct-current source is called upon to supply. Part of this power is delivered to the resonant circuit in the form of useful output, while the remainder is dissipated in heat at the plate of the tube. The plate loss at any instant is equal to the product of instantaneous plate current and instantaneous potential on the plate, and so varies according to the shaded area indicated in Fig. 124*f*, while the remainder represents useful output. Since it is desired that the useful output be as large as possible in proportion to the plate loss, it will be noted that the most efficient operating condition is when the plate current flows in the form of very short pulses that come when the instantaneous plate potential is small compared with the battery voltage.

A number of methods have been devised for direct experimental determination of plate loss. An obvious means is to dissipate the useful output of the oscillator in a resistance coupled to the tank circuit. The difference between the direct-current power supplied to the tube and the $I^2R$ loss measured in this resistance represents the plate losses of the tube plus the tank-circuit losses. The latter can be calculated by measuring the circulating current and the resistance of the tank circuit. The trouble with this method is that the efficiency of large oscillators is of the order of 70 to 80 per cent, so that the plate loss represents a small residual that cannot be determined very accurately by such an indirect method.

The most satisfactory experimental method involves the direct determination of the losses in the tube. In water-cooled tubes this can be accomplished by taking advantage of the fact that practically all of the energy dissipated in the tube is carried away in the cooling water, so that by measuring the rate of flow and the temperature rise of the cooling water one can calculate the energy dissipation within the tube.[1] The result thus obtained includes the filament heating power, which can be evaluated by direct measurement, and also the grid losses, which are relatively small compared with the plate loss.

In tubes having glass envelopes the total loss in the tube may be similarly obtained by inclosing the tube and cooling it with an air blast. The temperature rise and rate of flow of the cooling air then give dissipation in the tube as before. A modification of this method consists in immersing the tube in an oil bath and determining the rate of rise of temperature during operation. Several pyrometer methods have also been used to determine losses in the tube. The most successful of these measures the temperature of the glass envelope at a number of points by means of thermocouples.[2] The temperatures are recorded with the tube in normal operation, after which the grid exciting voltage is removed and the direct-current plate loss varied by adjusting the direct-current grid bias until on the average the same glass temperatures are obtained as in normal operation. The direct-current power required is then clearly the sum of grid, plate, and dielectric losses occurring in the tube during normal operation. In applying such a method it will usually be found that the temperature distribution during operating and non-operating conditions is somewhat different, and this is the principal source of error in the method.

The grid driving power at any instant is equal to the product of the instantaneous grid current and the grid exciting voltage,

[1] This method is described by A. Hoyt Taylor and H. F. Hastings, The Determination of Power in the Antenna at High Frequencies, *Proc. I.R.E.*, vol. 19, p. 1370, August, 1931. When the load is coupled to the tank circuit, the difference between the direct-current plate power and the plate loss obtained from the cooling water represents the energy converted to radio-frequency power. With the coupled load removed and the same circulating current as before in the tank circuit, the difference between these quantities now represents the total tank-circuit losses. The authors suggest that the coil losses can be segregated from the other tank-circuit losses by cooling the coils with water and determining the amount of heat carried away in this cooling water.

[2] See A. Crossley and R. M. Page, A New Method for Determining the Efficiency of Vacuum Tube Circuits, *Proc. I.R.E.*, vol. 16, p. 1375, October, 1928.

and so varies as shown in Fig. 124*g*. It will be noted that the grid current flows at or very near the crest of the exciting voltage, so that for all practical purposes the average grid driving power can be taken as the product of average grid current (*i.e.*, d-c grid current) times the crest value of the exciting voltage.[1] A portion of this grid driving power is dissipated in heat at the grid of the tube, while the rest, which is normally the largest part, is consumed in the grid bias. When the bias is derived from a leak-condenser combination, the $I^2R$ loss in the grid leak is supplied by the grid driving power. If the bias is obtained from a generator or batteries, the grid excitation delivers to the bias source an amount of energy equal to the product of d-c grid current and bias voltage.

The power loss at the grid of the tube at any instant is equal to the product of grid voltage and grid current at that instant, and so varies as shown by the shaded area in Fig. 124*g*. The average grid loss cannot be obtained with any degree of accuracy by taking the difference between the approximate grid driving power calculated from the d-c grid current as above, and power dissipated in the bias. Relatively little attention has been paid to experimental methods of measuring grid loss, and it appears that the most satisfactory means now available for evaluating this quantity is by use of oscillograms of grid voltage and current. It might be possible, however, to use an experimental technique involving the accurate determination of total driving power by a substitution method, and then subtracting from this the bias loss.

The emission of secondary electrons from the grid causes the heat generated at the grid to be greater than the heating resulting from the energy which the grid absorbs from the exciting voltage. Thus if a primary electron in striking the grid ejects a secondary electron that is drawn to the plate, the net contribution to the grid current is nil, and no energy is absorbed from the grid exciting voltage. The impact of the primary electron against the grid produces heating that is wrongfully assigned to the

[1] This was pointed out by the author on p. 234 of "Radio Engineering." Experimental verification showing that this method normally gives a result having an error of less than 10 per cent is given by H. P. Thomas, Determination of Grid Driving Power in Radio Frequency Power Amplifiers, *Proc. I.R.E.*, vol. 21, p. 1134, August, 1933.

plate. The consequence of this situation is that the power dissipated at any electrode losing secondary electrons is greater than that calculated on the basis of voltage and current oscillograms, and the loss at the electrode receiving the secondary electrons is correspondingly less than that indicated by voltage and current oscillograms.

**56. Modulated Amplifiers and Oscillators.**—The proper functioning of the modulation process is most satisfactorily determined by measuring the peak and trough modulation and the carrier amplitude as a function of modulating voltage when a *pure* sine wave of modulation is applied. Methods for doing this are described in Sec. 58.[1] The modulation is distortionless when the measurements indicate that the peak and trough modulations are the same and are proportional to the modulating voltage, and that the carrier amplitude remains constant irrespective of the degree of modulation. The amount of distortion may depend somewhat upon the modulating frequency, so that measurements must be made at a variety of modulating frequencies if a complete analysis is desired.

Distortion may arise from improper grid bias or excitation of the modulated tube, improper tank-circuit impedance, tank circuit that is so sharply resonant as to discriminate against side-band frequencies, variable load coupled to the tank circuit by the amplifier tube following the modulated tube, varying load placed upon the preceding tank circuit by the grid of the modulated tube, and overloading of the modulator. It is essential that modulation measurements be made when the tubes normally associated with the modulated tube are in operation since these tubes affect the linearity obtained.

While modulation measurements such as outlined above determine if distortion is being introduced, one must also consider the voltage, current, and power relations existing in the modulated tube, since these affect both the fidelity of the modulation and the power output obtained. Class C modulated amplifiers are essentially Class C amplifiers biased to approximately twice cut-off, with the modulation obtained by superimposing the

[1] The peak modulation represents the degree of modulation based upon the maximum amplitude of the wave envelope, while the trough modulation is the degree of modulation based upon the minimum amplitude of the wave envelope during the modulation cycle.

modulating voltage upon the direct-current plate potential. The voltage and current relations that exist at any part of the modulation cycle are hence similar to those that exist in an ordinary Class C amplifier and can be investigated with crest voltmeters as discussed in Sec. 54. It is to be noted that, if measurements are made when the amplifier is being modulated, the crest meters indicate the maximum positive potential reached by the grid during the modulation cycle, while the minimum plate potential $E_{\min}$ recorded is the smallest value reached during the modulation cycle. One can expect $E_{\min}$ to be least at the trough of the modulation cycle, while if the maximum positive grid potential varies at all during modulation cycle it will be greatest at the peak of the modulation. Modulated power oscillators have much the same voltage and current relations as modulated Class C amplifiers, and so are tested in the same way.

In grid-modulated amplifiers the minimum plate potential $E_{\min}$ varies greatly with the modulation, and the adjustment should be such that at the peak of modulation $E_{\min}$ has the proper value for Class C amplifier operation. This means that in the absence of modulation the minimum plate potential $E_{\min}$ will be greater than half the direct-current plate-supply potential. Measurements of $E_{\min}$ with a trough meter while modulation is taking place give the value which $E_{\min}$ has at the crest of the modulation cycle. Suppressor-grid modulation acts very much as control-grid modulation, in so far as the voltage and current relations at the plate are concerned, and can be handled in very much the same way.

In every system of modulation, the modulation process is equivalent to superimposing the modulating voltage upon some electrode. It is therefore possible to test the linearity by varying the direct-current potential of this electrode through the same range of values as is encountered during modulation, and noting the way in which the output varies. This procedure is practicable only with very small tubes, however, because the power dissipated at the plate of a modulated tube varies greatly during the modulation cycle, and serious damage may result by operating continuously for even a few seconds under conditions corresponding to the part of the modulation cycle giving maximum plate dissipation.

In plate-modulated amplifiers and oscillators the degree of modulation is to a high degree of accuracy equal to the crest amplitude of the modulating voltage applied to the plate divided by the direct-current plate potential. This fact can be used to estimate degree of modulation and is sometimes made use of to indicate the modulation for measuring or monitoring purposes.

It is possible to detect excessive distortion in most systems of modulation by noting the d-c plate current of the tube being modulated. If the modulation is linear, the plate current will be substantially constant, while a considerable variation as the modulation goes on and off indicates distortion.

**57. Linear Amplification of Modulated Waves.**—Linear or Class B amplifiers differ from Class C amplifiers in that the grid bias is approximately the cut-off value. When a radio-frequency wave is applied to the grid of such a tube, the plate current flows in pulses that approximate half sine waves having an amplitude very nearly proportional to the amplitude of the exciting voltage. The linear amplifier is said to be distortionless when the amplitude of the voltage developed across the plate tank circuit is exactly proportional to the amplitude of the grid exciting voltage. Distortion can be most readily detected by modulating the exciting voltage with a *pure* sine wave and then measuring the peak and trough modulations, the degree of modulation, and carrier amplitude of the output wave, exactly as in the case of modulated amplifiers discussed in the preceding section. If distortion is present, it may be due to improper grid bias, excessive exciting voltage, improper load impedance, or a variable load impedance placed upon the tank circuit by the grid of the succeeding tube.

When proper instruments for measuring the modulation of the output wave are not available, it is possible to obtain an idea as to whether or not the distortion is excessive by observing the d-c plate current of the linear amplifier tube as the modulation is applied. The current should be substantially constant.

The voltage and current relations in a linear amplifier must be such that at the peak of the modulation cycle the minimum plate voltage $E_{min}$ and the maximum instantaneous grid potential $E_+$ will have the proper values corresponding to normal Class C amplifier operation, *i.e.*, $E_{min}$ equal to or slightly more than $E_+$. In the absence of modulation the maximum grid potential is usually negative and $E_{min}$ is greater than one-half the direct-current plate voltage. Measurements of $E_+$ and $E_{min}$ made by crest meters while modulation is present give values that exist at the peak of the modulation cycle.

**58. Modulation Measurements.**—The important characteristics of a modulated wave are the average, maximum, and minimum amplitudes of the envelope. The average amplitude is equal to the carrier voltage, while the relation of the maximum

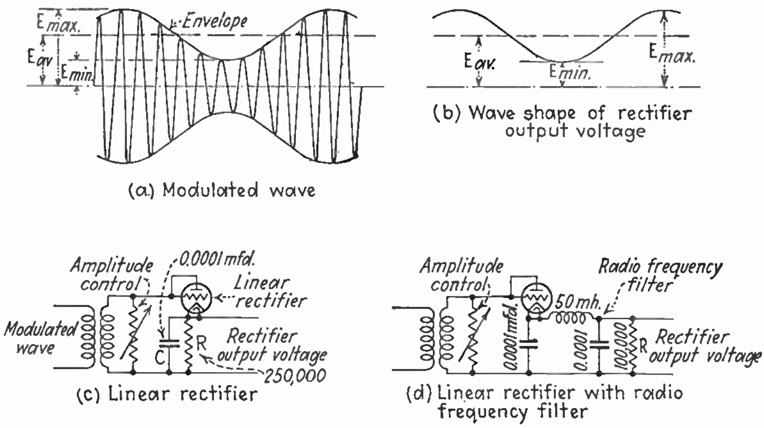



FIG. 128.—Linear rectifiers for developing a pulsating voltage that reproduces the modulation envelope.

and minimum amplitudes to the average give the peak and trough modulations according to the formulas

$$\text{Positive peak modulation} = \frac{E_{max} - E_{av}}{E_{av}} \tag{53a}$$

$$\left.\begin{matrix}\text{Negative peak or trough} \\ \text{modulation}\end{matrix}\right\} = \frac{E_{av} - E_{min}}{E_{av}} \tag{53b}$$

where $E_{max}$, $E_{min}$, and $E_{av}$ are respectively the maximum, minimum, and average envelope amplitudes as shown in Fig. 128.

The most satisfactory method of measuring these modulation characteristics is illustrated in Fig. 128 and consists of a linear rectifier which develops a pulsating output voltage that varies according to the wave envelope, together with means for determining the average, maximum, and minimum amplitudes of this pulsating voltage.[1] The rectifier can be a heater-type triode tube made to serve as a diode by tying the grid and plate together. The load resistance $R$ should be large compared with the diode resistance so that the voltage developed across $R$ by the rectified current may be a faithful reproduction of the modulation envelope. The shunting capacity $C$ should be small enough so that the ratio $X_c/R$ is at least unity at the highest modulation frequency at which measurements are to be made, and at the same time $C$ must be sufficiently large so that its reactance at the carrier frequency does not exceed a few per cent of the resistance $R$. Finally the carrier voltage applied to the diode should have an amplitude of at least 20 volts effective.[2] Suitable proportions are indicated in Fig. 128*c* for broadcast work. Where the relationship between the highest audio frequency in the modulation and the carrier frequency is such that one cannot make the capacity $C$ small enough to permit following the modulation envelope without making the reactance to the carrier frequency too large, then a simple filter such as shown in Fig. 128*d* can be employed.

The maximum and minimum envelope amplitudes can be measured by peak and trough meters such as described in Sec. 7, while the average amplitude can be determined from the direct current in the load resistance. A particularly convenient arrangement for making the necessary measurements upon the rectified wave is illustrated in Fig. 129 and has the advantage that it permits the direct determination of the peak and trough modulation.[3] In this method the dial of the slide-wire potentiometer is graduated into approximately 220 divisions which are

[1] See Balth. van der Pol and K. Posthumus, Telephone Transmitter Modulation Measured at the Receiving Station, *Exp. Wireless and Wireless Eng.*, vol. 4, p. 140, March, 1927.

[2] The requirements for the capacity $C$ arise from the fact that if the voltage across the condenser is to follow the modulation envelope the capacity must not be too large. See "Radio Engineering," p. 286. At the same time the capacity must be a virtual short circuit to the carrier frequency so that there will be no radio-frequency voltage appearing across the load resistance. The linearity of any detector is improved by using a large signal, and by making the load resistance high compared with the rectifier resistance.

[3] See W. N. Tuttle, Modulation Measurements on Broadcast Transmitters, *General Radio Experimenter*, vol. 5, p. 1, March, 1931.

proportional to resistance, and these calibrations are labeled as shown in Fig. 129*d* with the point −100 corresponding to zero voltage between slide and negative side of the battery. Thus for any setting of the potentiometer dial the voltage developed to ground is given by the dial reading as a percentage of the amount by which the voltage actually developed differs from the voltage at the point marked zero on the dial. In measuring the modulation of a wave the input to the linear detector is adjusted so that the direct-current voltage developed across the load

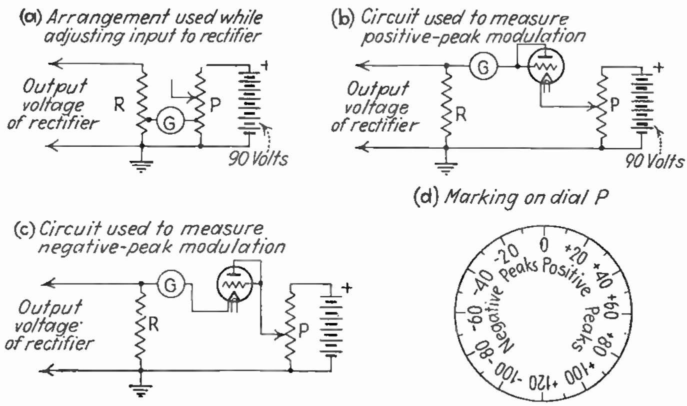



FIG. 129.—Details of arrangement for determining degree of modulation directly from the output voltage developed by a linear rectifier.

resistance $R$ is equal to the direct-current voltage corresponding to the point marked zero on the dial $P$. This is done by throwing a switch which arranges the connections as shown in Fig. 129*a*, and then adjusting the input to the linear rectifier until the galvanometer $G$ just begins to show current. The tap on the resistance $R$ which is used in this adjustment is so located that it develops about 5 per cent of the voltage across $R$, while the tap upon the potentiometer $P$ is likewise located so that it develops exactly this same percentage of the output voltage of the potentiometer when the potentiometer dial is set to read zero. This arrangement is employed so that a minimum effect will be produced by $P$ upon the alternating voltage developed across the resistance $R$. After the input has been adjusted in this way, the positive peak modulation is measured with a peak voltmeter as shown in Fig. 129*b*, by employing a heater-type triode with the grid and plate tied together to give diode action.

The setting of the dial $P$ at which the galvanometer $G$ just begins to show a trace of current gives the percentage modulation of the positive peak without further calculation. The negative peak modulation is measured in a similar manner by reversing the input to the peak voltmeter as shown in Fig. 129$c$, and adjusting the potentiometer until $G$ just shows a trace of current.

A modulation meter such as just described will measure degree of modulation to within a few per cent. The accuracy is highest when the modulation frequency is low and is also best when the degree of modulation is reasonably high.

*Miscellaneous Methods of Measuring Modulation.*—The modulation of a wave can be determined by a number of other methods. Thus the output of the linear rectifier can be amplified and observed with a magnetic oscillograph. This arrangement is sometimes employed to monitor broadcast transmitters but has the disadvantage that the magnetic oscillograph discriminates against the higher audio frequencies.

It is possible to estimate the degree of modulation by observing the increase in the power of the wave that results from modulation. The power of a modulated wave is proportional to $(1 + m^2/2)$, so that 50 per cent modulation corresponds to 12½ per cent increase in power, 100 per cent modulation to a 50 per cent increase in power, etc. The relative power can be observed by noting the deflection produced on a thermocouple galvanometer having a scale proportional to current squared. The objection to this method is that the accuracy is very poor with small degrees of modulation, and that any change in carrier amplitude accompanying modulation is erroneously attributed to the modulation.

The cathode-ray oscillograph can be used to determine the degree of modulation as discussed in Sec. 79 and is widely used to monitor broadcast transmitters. The cathode-ray oscillograph has the advantage of operating directly from the modulated wave so that the possibility of error is reduced to a minimum. At the same time it has the disadvantage that the width of the luminous spot limits the accuracy, and that one must have access to the modulating voltage in order to obtain a suitable sweep voltage.

A peak vacuum-tube voltmeter combined with a linear rectifier as shown in Fig. 130 can be used to measure the carrier amplitude and positive peak modulation. The peak voltmeter shown consists of an ordinary grid-leak grid-condenser type of power detector in which the leak-condenser time constant $RC$ is so high that the voltage developed across the condenser $C$ by the rectified grid current is not able to follow the modulation envelope but holds at substantially the peak value reached during the modulation cycle. The meter in the plate circuit of this detector tube then gives a deflection which is a measure of the maximum amplitude of the modulated envelope. The linear rectifier shown in Fig. 130 is a distortionless grid-leak grid-condenser power detector and differs from the peak instru-

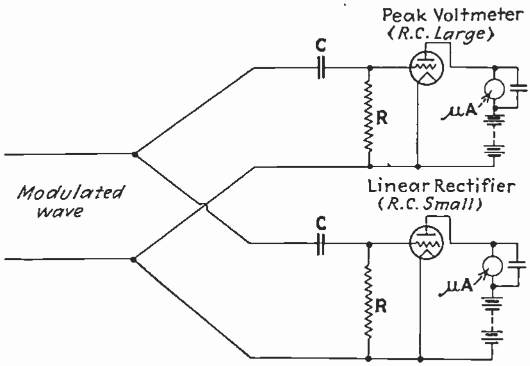



FIG. 130.—Method of obtaining continuous indication of positive peak modulation.

ment in that the time constant $RC$ of the leak-condenser combination is small enough to permit the voltage across the condenser $C$ to follow the modulation envelope. A direct-current meter in the plate circuit of this tube then gives an indication that is a measure of the average or carrier amplitude of the wave, so comparison of the two plate meter readings gives the positive peak modulation. This method has the merit of giving a continuous indication of the degree of modulation and so is especially useful for monitoring purposes.

Very small degrees of modulation, such as the hum produced by alternating filament current, can be determined by using an amplifier-detector type of vacuum-tube voltmeter to measure the alternating potential developed across the resistance $R$ of Fig. 128.

Waves having a known degree of modulation must sometimes be produced for the purpose of calibrating modulation meters or making laboratory measurements. The most convenient method of doing this is to set up a plate-modulated oscillator as shown in Fig. 131 and then determine experimentally the relationship between amplitude of oscillations and plate voltage as shown. From this curve, one can immediately determine the modulation envelope when a known alternating modulating voltage is superimposed upon the battery voltage as shown in Fig. 131. In order that the actual modulation may

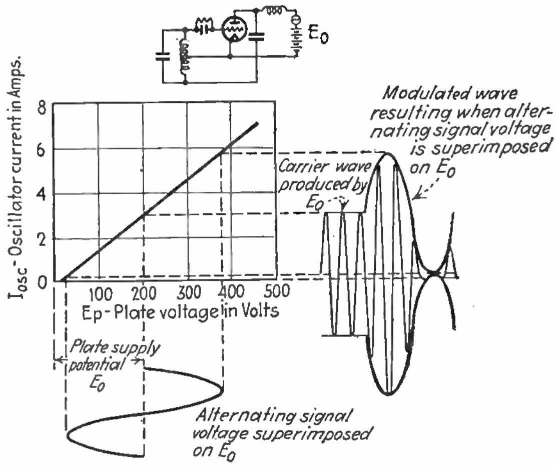



FIG. 131.—Plate-modulated oscillator, together with relation between amplitude of oscillations and plate voltage, showing how one may derive modulation envelope when a given alternating modulating voltage is superimposed upon the battery voltage.

follow a curve such as shown in Fig. 131, the time constant $RC$ of the grid leak-condenser combination must be small enough to permit the bias to follow the modulation, and the tuned circuit must not have such a high $Q$ as to discriminate against the side bands. The production of waves with known modulation is discussed further in Sec. 74.

*Frequency and Phase Modulation.*—Frequency and phase modulation are important because they often occur as undesired by-products of amplitude modulation. Frequency and phase modulation are essentially the same as far as the results are concerned although produced by different faults in the transmitter.

Frequency modulation accompanying amplitude modulation can be detected by modulating the carrier at a low frequency such as 60 cycles, and heterodyning the resulting modulated wave to give a beat note of about 500 cycles. The beat note modulated at 60 cycles is then recorded on an oscillograph. If the waves obtained are all equally spaced as shown in Fig. 132*a*, then no frequency modulation is present; but if there is a variable spacing as shown in Fig. 132*b*, then the frequency of the original

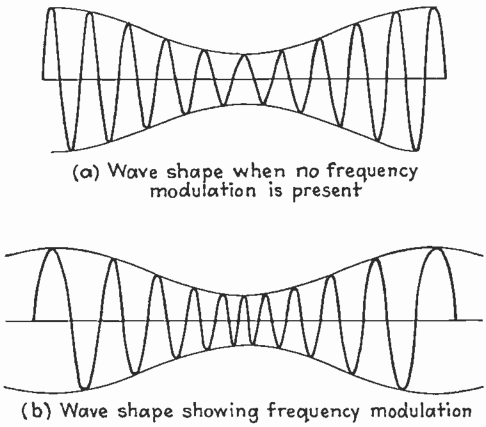



FIG. 132.—Beat notes obtained when carrier has (*a*) no frequency modulation, and (*b*) considerable frequency modulation.

carrier varies with the modulation.[1] Phase modulation can be determined with the same technique by integrating changes in the frequency which appear on the oscillogram.[2] This results from the fact that an advancing phase causes the effective frequency to be increased, so that an integration of frequency change with respect to time gives the phase shift.

The cathode-ray tube is very useful in investigating frequency and phase modulation in transmitting equipment, and its use for the purpose is discussed in Sec. 79.

[1] See Ralph Bown, DeLoss K. Martin, and Ralph K. Potter, Some Studies in Radio Broadcast Transmission, *Proc. I.R.E.*, vol. 14, p. 57, February, 1926.

[2] See A. Hund, "High-frequency Measurements," p. 369, McGraw-Hill Book Company, Inc., 1933.

# CHAPTER XI

## MEASUREMENTS ON RADIO WAVES, ANTENNAS, AND TRANSMISSION LINES

**59. Field Strength of Radio Waves.**—The field strength of a radio wave is determined by measuring the voltage which the wave induces in an antenna. The relationship between this

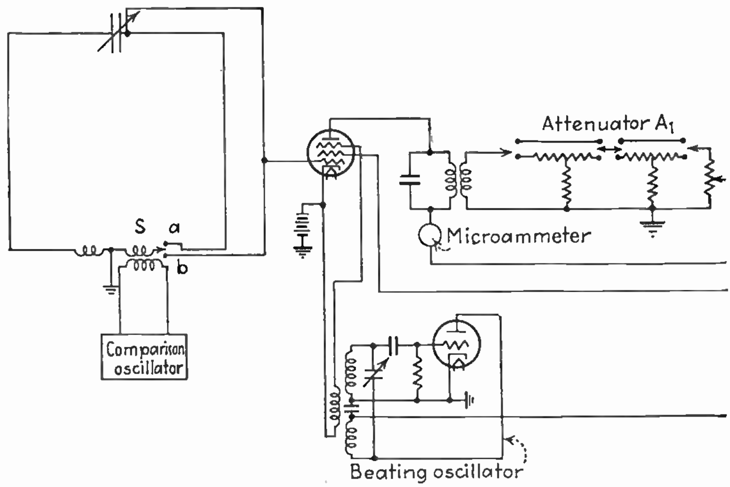



FIG. 133.—Typical field-strength equipment of

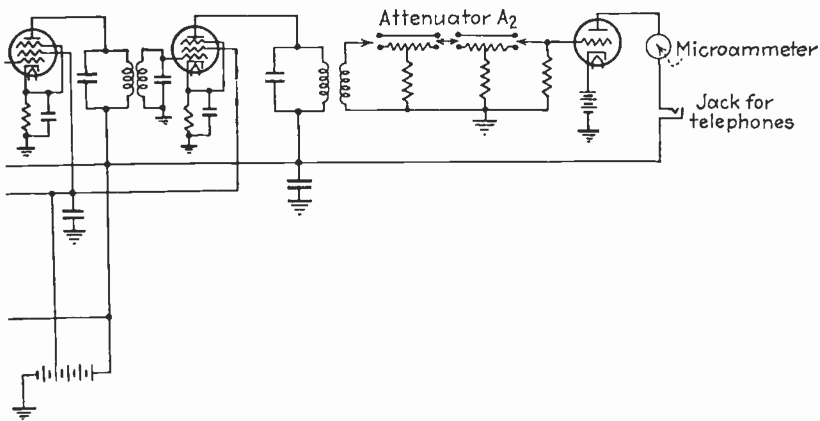
induced voltage and the field strength can be obtained by calculation in the case of loop and similar antenna systems, and by measuring the effective height of the antenna in other cases.

The usual method of measuring the voltage induced in an antenna is shown in Fig. 133,[1] and makes use of a superheterodyne receiver with an adjustable attenuator in the intermediate-frequency amplifier for adjusting the gain of the receiver in accurately known amounts. The first detector of the receiver

[1] For further details see H. T. Friis and E. Bruce, A Radio Field-strength Measuring System for Frequencies up to Forty Megacycles, *Proc. I.R.E.*, vol. 14, p. 507, August, 1926.

is adjusted so that the intermediate-frequency output is proportional to the signal voltage applied to the detector grid for signals up to several volts. The first detector is also arranged so that it can be made into a vacuum-tube voltmeter by turning off the beating oscillator. The comparison oscillator shown in the figure is a conventional oscillator provided with reasonably complete but not elaborate shielding, and an adjustable but uncalibrated output attenuator.

The measuring procedure is as follows: The signal is tuned in and the intermediate-frequency attenuator is adjusted to an attenuation $\alpha_1$ that gives a convenient deflection on a direct-current microammeter located in the plate circuit of the second detector.[1] The beating oscillator of the superheterodyne receiver is now turned off and the comparison oscillator set in operation at the frequency of the wave being measured. The




the intermediate-frequency attenuator type.

output of the comparison oscillator is coupled into the antenna system and adjusted until a voltage $E$ (commonly 1 volt) appears on the grid of the first detector as determined by using the first detector as a vacuum-tube voltmeter with a microammeter in its plate circuit. The beating oscillator is now turned on again, and the attenuator in the intermediate-frequency amplifier

[1] In the discussion to follow the attenuation $\alpha$ represents the multiplying factor which the attenuator introduces. Thus when $\alpha = 0.001$, it means that the attenuator reduces the voltage to 0.001 of the value with no attenuation.

readjusted until the output at the second detector is the same as that originally produced by the signal. If the attenuation required is $\alpha_2$ when the input voltage is $E$, then the input voltage required to produce the output when the attenuation is $\alpha_1$ is $E(\alpha_2/\alpha_1)$. This is the voltage which the signal produces at the grid of the first detector, and it now remains to determine the relationship between this voltage and the voltage actually induced in the antenna. This can be done by leaving the comparison oscillator untouched while coupling its output coil directly to the grid of the first detector by throwing switch $S$ to position $b$, and readjusting the attenuation to the value $\alpha_3$ that is required to give the standard output. The rise of voltage between the antenna and the detector grid is then obviously $\alpha_3/\alpha_2$, so that the voltage actually induced by the signal in the antenna is hence $E\alpha_2^2/\alpha_1\alpha_3$.

*Design Considerations.*—The receiver of Fig. 133 must be designed so that the beating oscillator is not affected by tuning the input circuit of the grid of the first detector. When a triode first detector is employed, this can be accomplished by arranging the beating oscillator to plate modulate the detector, and at the same time employing a neutralized circuit to prevent coupling between grid and plate. An alternative arrangement is to use a pentode first detector with suppressor grid modulation.

The linear relation required between intermediate-frequency output and signal at the grid of the first detector can be satisfied by operating the tube as a full-wave square-law device. The first detector tube can be used as a vacuum-tube voltmeter without any change in connections by merely turning off the beating oscillator and making use of the plate-circuit microammeter.

The attenuator in the intermediate-frequency amplifier can consist of T sections, each of which reduces the voltage amplification by a factor of $\frac{1}{10}$ or $\frac{1}{100}$, supplemented by a slide-wire potentiometer to give continuous adjustment. The best impedance for the attenuator sections is of the order of 1000 ohms, and care must be taken to insure that the load actually present is a resistance of the correct value so that switching attenuator sections in and out will not upset the impedance relations. For this reason the attenuation should be introduced as shown in Fig. 133, by using a transformer to effect the best possible transfer of energy from the plate of the tube to the attenuator, but using a resistance load rather than a transformer in the output. With an attenuator impedance of 1000 ohms means that the amplification of the stage when the attenuation is switched out is not large, and at least one conventional intermediate-frequency stage must be depended upon to develop adequate amplification. To prevent excessive differences in power level, attenuating and amplifying stages should alternate with each other as in Fig. 133 except when the total attenuation required is small (measurements to be made only on strong signals), when all the attenuation can be placed in the plate circuit of the first detector. When the attenuation is divided, a disproportionate fraction should still be located in the plate circuit of the first detector to prevent the possibility of overloading in the intermediate-frequency amplifier.[1] The attenuation resistances can be wound on thin cards by using a type of winding having low capacitive and inductive effects as described in Sec. 23, and must be carefully mounted and shielded. In order to minimize attenuator troubles the intermediate frequency should not exceed 300 kc.

The shielding requirements are not severe since the only possible sources of trouble are regenerative couplings in the intermediate-frequency amplifier that by-pass the attenuator networks, and coupling between the comparison oscillator and the antenna system. The latter coupling can be readily handled because the comparison oscillator is not operated when weak signals are being received, and the shielding need only be sufficient to prevent stray couplings from introducing voltages into the antenna system which are appreciable compared with those which the comparison oscillator induces in producing a potential of $E$ volts at the grid of the first detector.

The details shown in Fig. 133 may be varied in numerous respects when desired. In particular a vertical antenna can be used by making obvious modifications. Additional selectivity may also be added between the antenna and first detector when necessary. Such selectivity should preferably consist only of

[1] Thus, in Fig. 133, overloading cannot occur if the attenuation of $A_2$ is such that with maximum attenuation at $A_2$ the voltage at the grid of the second detector is less than the voltage at the grid of the tube just before attenuator $A_2$.

additional tuned circuits without amplification in order to avoid pick-up troubles from the comparison oscillator.

*Substitution Method of Measuring Field Strength.*—This method is illustrated in Fig. 134 and makes use of a standard signal generator and a sensitive radio receiver.[1] The measuring procedure consists in tuning in the signal on a radio receiver, rotating the loop for maximum response, and adjusting the volume until a convenient deflection is obtained upon a direct-current microammeter placed in the plate circuit of the detector. The

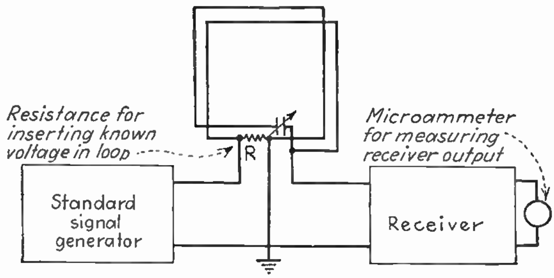



FIG. 134.—Method of measuring field strength with the aid of a standard signal generator and an ordinary receiver.

loop is then rotated so that no signal is induced, and a known voltage of the same frequency as the signal, obtained from a standard signal generator, is introduced into the antenna as a drop across the resistance $R$ of 1 or 2 ohms, and adjusted in amplitude until the radio receiver indicates the same output as produced by the actual signal. The known voltage is then obviously equal to that induced by the passing wave. This method of measurement has the disadvantage that the signal generator must be thoroughly shielded from the radio receiver and must be capable of producing accurately known small voltages. It is hence best adapted to long-wave measurements, where these requirements are most readily met, but can be used at broadcast frequencies with fair success. The method has the further disadvantage that when it is used with a vertical antenna one must wait for the transmitter to stop sending before introducing the signal-generator voltage.

[1] For further information see Axel F. Jensen, Portable Receiving Sets for Measuring Field Strengths at Broadcasting Frequencies, *Proc. I.R.E.*, vol. 14, p. 333, June, 1926.

*Miscellaneous Considerations.*—A loop antenna is normally used in making field-strength measurements because, when the precautions discussed in Sec. 60 are taken, the relationship between the field strength and the resulting induced voltage acting around the loop can be accurately determined by calculation. The loop also has the advantage of being portable. When a loop antenna is employed, it is possible to make measurements of field strength down to about 10 microvolts per meter by using either of the methods that have been described. A vertical antenna can be used in evaluating weaker fields but has the disadvantage that the effective height of such an antenna for each particular installation must be determined by experiment, with the result that the vertical antenna is suitable only for permanent installations.

In using field-strength measuring equipment such as has been described, some thought must be given as to the components of the passing wave that induce voltage in the measuring antenna. Thus a loop with its plane vertical responds not only to the horizontally traveling vertically polarized component of the wave, but also to any downward-traveling component that may be present. The actual response obtained is then the resultant of these two individual effects. Considerations similar to these apply to most types of antennas.

Fading signals such as are often encountered at broadcast and higher frequencies present a rather difficult measuring problem. A method commonly employed to handle this situation is to determine the field strength at the peaks of the fading cycle. Successive peaks are normally of substantially the same amplitude because they represent moments when the various wave components reaching the receiver all add. Another possibility is to use an output meter that gives an averaging effect, such as a thermocouple instrument, or a galvanometer with a long period, in which case the measurements indicate the average field strength over a time interval.

*Standard Field Generators.*—The difficulties of obtaining accuracy at frequencies of the order of 60 megacycles and higher are such that the most satisfactory means of obtaining a known calibrating voltage of such a high frequency is to produce at the measuring antenna itself a known field from a local source which can be termed a "standard field generator." This standard field generator is a compact portable oscillator with a small loop antenna provided with a thermocouple to measure the loop current. The field produced in the vicinity of such an arrangement can be calculated with good accuracy from the dimensions involved and from the loop current. It is thus possible to compare the unknown field with the field from the standard field generator, and to calibrate the intermediate-frequency attenuator type of measuring equipment for ultra-high frequencies.[1]

*Continuous Recording of Field Strength.*—Fading and diurnal changes in field strength make continuous records of signal strength necessary if a really complete picture is to be obtained of the signals from a distant transmitter. All such systems that have been developed make use of a radio receiver provided with some form of automatic volume control, with which is associated a continuous recorder. Arrangements that have been employed include using the current flowing through the bias resistor of an ordinary automatic volume-control system to operate the recorder;[2] using the recorder to indicate the output of the radio receiver, and obtaining the automatic volume control by connecting the recorder shaft to the volume control of the receiver in such a way that as the output increases the receiver gain is decreased, and vice versa;[3] and, finally, using arrangements in which provision is made so that if the output over a period of 5 or 10 sec. differs from a standard value the receiver gain is increased or decreased as required by a definite amount, with the recorder operating from the volume-control shaft.[4]

[1] For further information on the use of standard field generators in the measurement of field strength, and in particular for a discussion of the problems involved in avoiding errors as a result of ground reflections making the field differ from the calculated value, see: J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, Ultra-short-wave Propagation, *Proc. I.R.E.*, vol. 21, p. 427, March, 1933, and B. Trevor and R. W. George, Notes on Propagation at a Wavelength of Seventy-three Centimeters, *Proc. I.R.E.*, vol. 23, p. 461, May, 1935.

[2] G. D. Robinson, Wide Range Scales for Fading Records by Electrical Means, *Proc. I.R.E.*, vol. 19, p. 247, February, 1931.

[3] P. A. deMars, G. W. Kenrick, and G. W. Pickard, Use of Automatic Recording Equipment in Radio Transmission Research, *Proc. I.R.E.*, vol. 19, p. 1618, September, 1931.

[4] W. W. Mutch, A Note on an Automatic Field Strength and Static Recorder, *Proc. I.R.E.*, vol. 20, p. 1914, December, 1932.

Continuous recorders of field strength must be capable of handling a wide range of signal strength, which makes it desirable that the indication be approximately proportional to the logarithm of the field strength. The receiver employed must be stable in its characteristics and should have the voltage of its power supply closely regulated. Calibration can be readily

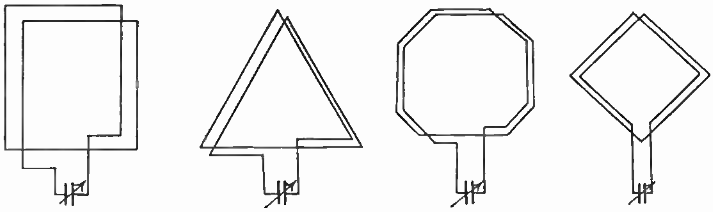


Fig. 135.—Examples of typical loop antennas.

made by periodically introducing known voltages in the antenna system from a signal generator.

**60. Loop Aerials.**—The importance of the loop aerial arises as a result of the fact that it is directional, is independent of the ground as far as reception of vertically polarized waves is concerned, is portable, and that its characteristics can be accurately determined by calculation. A loop antenna is essentially a large coil of any convenient section (see Fig. 135 for examples). Such an aerial abstracts energy from passing waves as a result of phase differences between the voltages induced in the opposite legs. Thus consider the case of a rectangular loop in the path

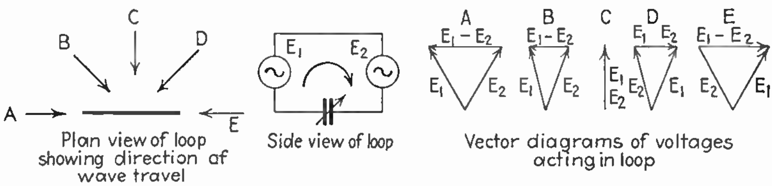



Fig. 136.—Vector diagram showing how the voltages induced in the two sides of a loop by a passing radio wave combine to give a resultant voltage acting around the loop.

of a horizontally traveling vertically polarized radio wave. When the plane of the loop is perpendicular to the direction of travel, the voltages induced in the two vertical legs are equal and in the same phase but, being directed around the loop in opposite sense, cancel each other and result in zero response. As the plane of the loop is brought nearer to parallel with the direction of wave travel, the wave front reaches the two vertical legs at slightly different times. This causes a phase difference between the voltages induced in these two legs, giving rise to a resultant voltage that circulates current around the loop and is maximum when the plane of the loop is parallel to the direction of wave travel. Vector diagrams illustrating the situation for several loop orientations are shown in Fig. 136, and the resulting directional characteristic in a horizontal plane is as in Fig. 137.

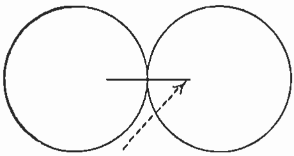



FIG. 137.—Directional characteristic of a loop antenna. This applies to loops of all shapes.

The resultant voltage acting around a rectangular loop is given by the following equation:[1]

$$\text{Resultant voltage acting around loop} = 2\epsilon l N \sin\left(\frac{\pi s}{\lambda}\cos\theta\right) \tag{54}$$

where

$\epsilon$ = strength of radio wave in volts per meter
$l$ = height of loop in meters
$s$ = width of loop in meters
$N$ = number of turns in loop
$\lambda$ = wave length of radio wave in meters
$\theta$ = direction of travel of wave with respect to plane of loop.

In practical loops the size is small compared with a wave length, so that $\sin(\pi s/\lambda \cos\theta)$ may be written as $\pi s/\lambda \cos\theta$ without appreciable error, giving

$$\text{Resultant voltage acting around the loop} = 2\pi\epsilon N\frac{ls}{\lambda}\cos\theta \tag{54a}$$

$$= 2\pi\epsilon N\frac{(\text{loop area})}{\lambda}\cos\theta \tag{54b}$$

Equation (54*b*) applies to loops of all shapes provided only that the loop is small compared with a wave length.

*Unbalances to Ground in Loop Antennas.*—The loop antenna behaves as described above if balanced with respect to voltages

[1] This formula can be readily derived as follows: The voltage induced in each vertical leg is $\epsilon Nl$, while the phase difference between the voltage is $2\pi s/\lambda \cos\theta$ radians, since the wave front must travel a distance $s\cos\theta$ to pass from one leg to the other. Subtracting the voltages in the two legs while taking into account this phase difference gives Eq. (54).

and capacities to ground. The effect of unbalances to ground can be explained with reference to Fig. 138*a*, which shows the loop unbalanced to ground as a result of the voltage developed across the tuning condenser. Current flows through $C_g$ to ground and returns up through the left-hand side of the loop as a result of the voltage developed across this leg, but this current is not the same as that which flows through $C_g'$ because of the reactance of the tuning condenser which is in the circuit, and because of the voltage developed across the tuning condenser by the loop current. These unequal currents in the two vertical legs produce a current that circulates around the loop just as does the

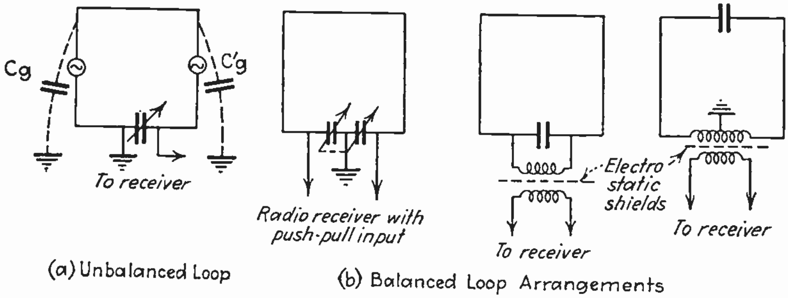



FIG. 138.—Unbalanced and balanced loop arrangements.

desired pick-up, and which hence alters the characteristics from those obtained by calculation from Eq. (54).

This trouble can be avoided by balancing the loop with respect to ground as shown in Fig. 138*b*. In each of these balanced arrangements the two sides of the loop antenna are symmetrical with respect to ground so that the currents flowing to ground are equal and, being in opposite direction with respect to the normal loop circuit, produce no currents that circulate around the loop.

While loop antennas are normally thought of as operating from the vertically polarized components of horizontally traveling waves, down-coming waves (sky waves) are able to induce voltages in the top and bottom sides of the loop. The magnitude of this effect depends upon the height of the loop above ground and upon the ground characteristics. When the loop is used under conditions where this action causes trouble, one must replace the loop with a different type of antenna, such as an Adcock antenna (see Sec. 61) or a plain vertical antenna.

**61. Direction of Wave Travel.**[1]—The direction in which a wave is traveling can be determined by the use of a receiver in conjunction with a loop antenna. The procedure is to rotate the loop until zero signal is obtained, for which condition the plane of the loop is perpendicular to the direction of wave travel. The position of the minimum rather than the maximum response is used since it permits much more accurate settings.

In order to determine from which side of the loop the waves are arriving, a vertical antenna is used in conjunction with the loop.[2] The vector diagrams of Fig. 136 show that when the loop is adjusted for maximum response the resulting voltage acting around the loop is 90° out of phase with the wave at the center of the loop, and may either lead or lag according to the direction from which the wave arrives. A vertical antenna located at the loop center and coupled inductively to the loop circuit will induce a voltage in the loop that is 90° out of phase with the current in the vertical antenna. It will hence either add to or subtract from the voltage acting around the loop, depending on the direction of arrival of the waves. The procedure for obtaining the sense of the bearing is first to adjust the loop to zero response with the vertical antenna disconnected. The loop is then rotated 90° in a specified direction in order to give maximum response, after which the vertical antenna is tuned to resonance and coupled to the loop circuit. The 180° uncertainty in the loop bearing is then removed by noting whether the presence of the vertical antenna causes the loop response to increase or decrease.

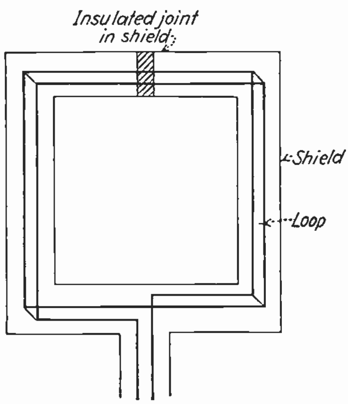


FIG. 139.—Electrostatically shielded loop antenna.

[1] For further information see R. L. Smith-Rose, Radio Direction Finding by Transmission and Reception, *Proc. I.R.E.*, vol. 17, p. 425, March, 1929; Frederick A. Kolster and Francis W. Dunmore, The Radio Direction Finder and Its Application to Navigation, Bur. Sta., *Sci. Paper* 245.

[2] For further details see Kolster and Dunmore, *op. cit.*

The loop must be carefully balanced to ground as shown in Fig. 138 and is also often shielded electrostatically by inclosing it in a pipe that is prevented from acting as a short-circuited turn by the insertion of an insulating bushing as shown in Fig. 139. Such an electrostatic shield makes the ground capacities constant for all positions of the loop and so prevents spurious circulating currents.

When the balanced loop is of the type having a ground connection, the currents flowing in this ground wire correspond to the currents flowing in a vertical antenna located at the loop center, and are often used to determine the sense of the bearing.

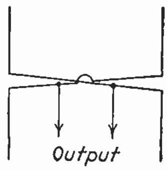



FIG. 140.—Simple form of Adcock antenna.

The bearings obtained by directional measurements upon a radio wave are often affected by near-by objects such as telephone and power wires, buried pipes, trees, buildings, which absorb energy from the wave and then reradiate it, or which induce voltages in the loop by direct induction. It is hence necessary to make an experimental calibration for each set-up if the highest accuracy is required.

The accuracy of bearings obtained with a loop antenna is also affected by down-coming (or sky) waves since these induce voltages in the horizontal members of the loop and cause a resulting current to flow in the loop even at what should be null position. The result is either a false bearing or a blurred minimum. Hence a loop antenna will give the direction of travel of a radio wave only when the sky wave is weak compared with the ground wave.

The errors in bearing caused by down-coming waves can be eliminated by replacing the loop with the Adcock antenna system, which in its simplest form consists of two spaced antennas connected as shown in Fig. 140. The action of such an antenna as far as vertically polarized horizontally traveling waves are concerned is similar to the loop, since the resultant current in the output coil of the Adcock antenna is proportional to the vector difference of the voltages induced in the two vertical members, exactly as is the case with the loop. The vertical component of a sky wave does not affect the Adcock antenna, however, since the connections are arranged so that all voltages induced in the horizontal members balance out. The result is that the bearings obtained depend only on the horizontally traveling component of the sky wave even when considerable down-coming component is also present. A carefully balanced and electrostatically shielded Adcock antenna will hence give accurate bearings at frequencies in excess of 5000 kc[1] where loops are useless because of the large sky wave and total absence of a ground wave.

The loop antenna when accurately calibrated will give bearings on waves from near-by radio transmitters that are accurate to within 1 or 2°. The accuracy diminishes, however, as the distance to the transmitter and the transmitter frequency increases because of the increased ratio of sky wave to ground wave. The distance from the transmitter over which satisfactory bearings can be obtained is greater by day than at night for the same reason. The maximum distance is of the order of 50 to 200 miles at a frequency of 500 kc, and may be as great as several thousand miles at a frequency of 20 kc. When the Adcock antenna system is used, accurate bearings can be obtained under practically all conditions provided the frequency is not too high.

**62. Static.**—Static presents a rather difficult measuring problem because of the impossibility of evaluating the disturbing effect of the random static noises in objective terms. When the static is a continuous rumble as is commonly the case at long waves, the disturbing effect can be measured in terms of the field strength of a just audible warbler tone or the field strength of just audible telegraphic signals. The average energy of the static in such cases is also indicative of the static intensity and can be determined by using in the field-strength measuring equipment some form of output system which averages the response.[2]

Observations upon individual static impulses can be obtained by the use of a cathode-ray tube. Methods have been devised whereby the direction from which a particular static impulse arrives can be determined with the aid of a cathode-ray tube,

[1] For a discussion of short-wave direction finding with the Adcock aerial see R. H. Barfield, Recent Developments in Direction Finding Apparatus, *Exp. Wireless and Wireless Eng.*, vol. 7, p. 262, May, 1930.

[2] A continuous recorder of static which gives indications proportional to the static power absorbed over a short-time interval is described by H. T. Friis, A Static Recorder, *Bell System Tech. Jour.*, vol. 5, p. 282, April, 1926.

and also means by which the wave form of individual impulses can be observed. This is described further in Sec. 79.

**63. Investigations of the Received Waves.**—The wave which reaches a receiving point usually consists of a ground wave with one or more sky-wave components superimposed. The ground wave travels along the surface of the earth with a slight forward tilt as shown in Fig. 141 and is vertically polarized at the earth's surface. The ground wave attenuates rapidly with distance at frequencies above 2000 kc but at broadcast frequencies is depended upon to produce the signal in the service area, and at still lower frequencies is of importance up to 1000 miles or more.

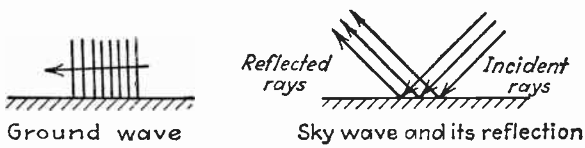



FIG. 141.—Character of ground wave, showing forward tilt of wave front, and also the reflection at the earth's surface of a sky wave. In the latter case the actual field is the resultant of the incident and reflected waves.

A sky wave consists of a ray which has been refracted back to earth by the ionized region in the upper atmosphere. Upon striking the surface of the earth this ray suffers reflection as shown in Fig. 141, so that the field actually observed at the receiving point is the resultant of the incident and reflected rays and so depends upon the electrical constants of the ground and the height above earth. The composition of the sky wave is complicated by the fact that the wave normally contains both vertically and horizontally polarized components which follow different laws of reflection at the earth's surface, and which fade independently. When several sky waves are present, each has its own angle of incidence, polarization, independent fading, etc.

The wave actually observed at a receiving point is the resultant of the various component waves present and can be investigated by the use of antennas and combinations of antennas arranged so that only certain components of the resultant wave produce a response. Thus a rod antenna will not respond when the rod is either parallel to the magnetic flux of the wave or at right angles to the wave front, since in both instances no flux cuts across the antenna.[1] Likewise a loop antenna when arranged in a vertical plane that is at right angles to the direction of wave travel will respond only to the resultant wave formed by the down-coming component of the sky wave and its ground reflection. Combinations of similar antennas with different orientations, or combinations involving different types of antennas, are also commonly used to give further information. Thus the polarization of a single sky wave in the presence of a ground wave can be determined by varying the transmitting frequency and observing the response on three loop antennas having planes oriented respectively $-45, 0$, and $+45°$ with the direction of the arriving waves.[2]

*Ground Characteristics.*—The ground is of utmost importance in connection with measurements on radio waves because of its action in reflecting sky waves and absorbing energy from the ground wave.[3]

The properties of the ground which are of importance are the dielectric constant $\epsilon$ and the conductivity $\sigma$. These quantities can be most readily determined by measurements upon soil samples. The procedure is to remove the soil with a suitable sampling cup and utilize it as the dielectric of a condenser. A substitution type of measurement is then employed to determine the conductivity and dielectric constant of the soil. The variability of soil makes it necessary to average the results obtained from numerous samples if the constants obtained are to have any significance.

**64. Kennelly-Heaviside Layer Studies.**—The ionized region in the upper atmosphere (called the Kennelly-Heaviside layer or ionosphere) is of such great importance in the propagation of radio waves that it is the subject of intensive investigation. One of the most effective means of investigating the ionosphere is the pulse method originated by Breit and Tuve.[4] In this method

[1] Use of such an antenna to determine the polarization of waves is described by Greenleaf W. Pickard, The Polarization of Radio Waves, *Proc. I.R.E.*, vol. 14, p. 205, April, 1926.

[2] This method was devised by Appleton and Ratcliffe and is described by A. L. Green, The Polarization of Sky Waves in the Southern Hemisphere, *Proc. I.R.E.*, vol. 22, p. 324, March, 1934.

[3] An excellent discussion of the effect of the ground upon radio waves is given by C. B. Feldman, The Optical Behavior of the Ground for Short Radio Waves, *Proc. I.R.E.*, vol. 21, p. 764, June, 1933.

[4] See G. Breit and M. Tuve, A Test of the Existence of the Conducting Layer, *Phys. Rev.*, vol. 28, p. 554, September, 1926.

short wave trains lasting perhaps $10^{-4}$ sec. are transmitted[1] and recorded on an oscillograph operated from the output of a radio receiver located within the range of the ground wave (often located in a room adjacent to the transmitter but shielded sufficiently to prevent excessive direct pick-up) and with the receiving antenna orientated for minimum pick-up from the transmitting antenna. The type of oscillogram obtained is indicated in Fig. 142. The first received pulse represents the

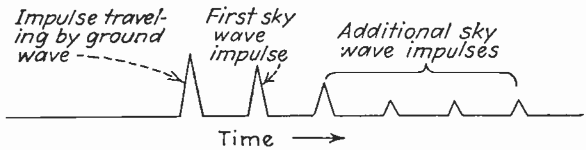



FIG. 142.—Typical oscillogram of signal received when a transmitter within ground-wave range sends out a short wave train.

ground wave, while the second represents a wave that has traveled up to the Kennelly-Heaviside layer and then been refracted back to earth. The time interval between the first two pulses is a measure of the difference in path lengths and can be used to obtain the height of the layer. The additional pulses, when present, represent other sky waves corresponding either to multiple reflections between earth and the ionized region or to refractions from higher layers. In interpreting the results given by pulse records it must be remembered that the wave travels along a curved path while in the ionized region, as shown in Fig. 143, and that the group velocity in the ionized region is less than the velocity of light. The height calculated on the basis of the triangular path $TAR$, by using the observed time delay and the velocity of light, is referred to as a "virtual

[1] Such pulses can be produced in a number of ways. Perhaps the most satisfactory method is by the use of a gas triode tube (Western Electric 269-A or RCA-885) to generate pulses of controllable frequency and duration which can then be used to modulate the transmitter. See J. P. Schafer and W. M. Goodall, Kennelly-Heaviside Layer Studies Employing a Rapid Method of Virtual Height Determination, *Proc. I.R.E.*, vol. 20, p. 1131, July, 1932.

A very simple system of producing pulse signals is to use as the transmitter an ordinary oscillator in which the grid-leak resistance is so high that interrupted oscillations are produced which have the desired pulse characteristics.

height" and is the thing actually determined. The true height can only be deduced approximately by analysis of virtual-height results obtained under various conditions.

Recent developments in the pulse method of investigation include means for producing continuous records which give virtual height as a function of time for each returned wave, and also means of obtaining the virtual height as a function of frequency. Methods have also been developed by which the time delays of the various received pulses can be observed visually by using a cathode-ray tube. Small changes in virtual height have been investigated by noting shifts in the phase of the refracted wave.[1]

The interpretation of these pulse records leads into fields which are far beyond the scope of this work, and one who is interested in pursuing the subject should consult the extensive literature that is available. In general the analysis is very difficult because the ionized region frequently possesses a stratified character. As yet a complete determination of actual electron density as a function of height has never been obtained for even a single case.[2]

The pulse method is limited to moderately high frequencies since at low frequencies the pulses are too short compared with the time of a cycle to give a satisfactory wave train, while at extremely high frequencies the skip distance is much greater than the range of the ground wave.

[1] T. R. Gilliland and G. W. Kenrick, Preliminary Note on an Automatic Recorder Giving a Continuous Height Record of the Kennelly-Heaviside Layer, *Proc. I.R.E.*, vol. 20, p. 540, March, 1932; H. R. Mimno and P. H. Wang, Continuous Kennelly-Heaviside Layer Records of a Solar Eclipse, *Proc. I.R.E.*, vol. 21, p. 529, April, 1933; J. P. Schafer and W. M. Goodall, Kennelly-Heaviside Layer Studies Employing a Rapid Method of Virtual Height Determination, *Proc. I.R.E.*, vol. 20, p. 1131, July, 1932; Lal C. Verman, S. T. Char, and Aijaz Mohammed, Continuous Recording of Retardation and Intensity of Echoes from the Ionosphere, *Proc. I.R.E.*, vol. 22, p. 906, July, 1934; T. R. Gilliland, Note on a Multifrequency Automatic Recorder of Ionosphere Heights, *Proc. I.R.E.*, vol. 22, p. 236, February, 1934; L. R. Hafstad and M. A. Tuve, An Echo Interference Method for the Study of Radio Wave Paths, *Proc. I.R.E.*, vol. 17, p. 1786, October, 1929.

[2] An excellent summary of the subject with numerous references to the literature is given by S. S. Kirby, L. V. Berkner, and D. M. Stuart, Studies of the Ionosphere and Their Application to Radio Transmission, *Proc. I.R.E.*, vol. 22, p. 481, April, 1934.

Appleton and Barnett[1] have investigated the Kennelly-Heaviside layer by varying the transmitted frequency and observing the variations that occur in the signal strength at a point which receives energy from both sky and ground waves. Because of the different path lengths of sky and ground waves the two waves will alternately add and subtract as the frequency is varied, and the increment in frequency required to change the relative phases by 180° can be used to estimate the layer

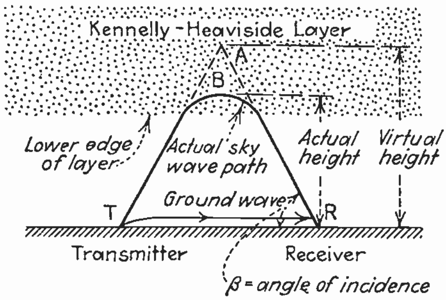



FIG. 143.—Diagram showing relation of actual and virtual height of a ray refracted by the Kennelly-Heaviside layer.

height. Theoretical analysis indicates that the layer height obtained in this way is the virtual height given by the pulse method of Breit and Tuve. This method of determining the height of the Kennelly-Heaviside layer can be used at low radio frequencies, but is not entirely satisfactory at high frequencies because of complications introduced by fading and multiple sky waves.

Appleton and Barnett have also made use of the angle of incidence of the down-coming sky wave to determine the height of the Kennelly-Heaviside layer. It is apparent from examination of Fig. 143 that, if the angle of incidence with which the sky wave strikes the earth and the distance between receiver and transmitter are known, then the virtual height of the Kennelly-Heaviside layer can be found by simple triangulation.

Still another way of determining the height of the Kennelly-Heaviside layer consists of observing the variations in field intensity as the distance between transmitter and receiver is

[1] See E. V. Appleton, Some Notes on Wireless Methods of Investigating the Electrical Structure of the Upper Atmosphere, *Proc. Phys. Soc.*, vol. 41, pt. II, p. 43, December, 1928.

varied.[1] Thus Hollingworth has found that at moderate distances from long-wave transmitters the signal strength alternately decreases and increases as the distance between transmitter and receiver is varied. This action is the result of alternate reinforcement and cancellation between sky and ground waves.

*Investigation of High-frequency Signals from Distant Transmitters.*—High-frequency signals arriving from distant transmitters are normally a combination of two or more sky waves which travel along different paths in reaching the receiver, arrive at different angles of elevation, and have polarizations, fading cycles, etc., that are quite independent. The only successful means that has been devised for investigating the character of this complex received wave is the pulse method, since each received pulse corresponds to a particular path and so can be studied by itself. In this way it is possible to determine the angle of incidence, the fading characteristics, angle of polarization, etc., separately for each path.[2]

*Selective Fading.*—When a signal reaches a receiving point along two or more different paths, the field strength that results is the vector sum of the various components. The signal strength is then very critical with respect to frequency because it takes only a very slight change in frequency to alter the difference in path lengths by one-half wave length and thereby to change completely the way in which the individual waves combine. At frequencies such as 15 mc it is thus not unusual for frequencies differing by as little as 100 cycles to give quite different signal strengths at a distant receiver, and the same effect is encountered at broadcast frequencies to a slightly lesser degree when a sky wave and ground wave combine. Inasmuch as voice-modulated signals contain a number of frequencies, considerable distortion is introduced as a result of the unequal transmission of the different side-band frequencies. This effect is termed *selective fading* and is most readily studied by varying or "wobbling" the frequency of an unmodulated carrier wave and observing the varia-

[1] See J. Hollingworth, Propagation of Radio Waves, *Jour. I.E.E. (London)*, vol. 64, p. 579, May, 1926.

[2] See T. L. Eckersley, Multiple Signals in Short-wave Transmission, *Proc. I.R.E.*, vol. 18, p. 106, January, 1930; H. T. Friis, C. B. Feldman, and W. M. Sharpless, The Determination of the Direction of Arrival of Short Radio Waves, *Proc. I.R.E.*, vol. 22, p. 47, January, 1934.

tion in signal strength produced at the receiver. Another general method of investigation involves modulating a fixed carrier frequency and investigating the received wave.[1]

**65. Antenna Measurements.** *Antenna Impedance.*—The impedance that an antenna offers does not vary with frequency in the same way as does the impedance of an ordinary resonant circuit because the inductance and capacity of an antenna are distributed rather than lumped. As a result of the distribution of the antenna constants, the impedance and power factor vary cyclically with increasing frequency as shown in Fig. 144. It will be noted particularly that the reactance alternates systematically between inductive and capacitive. These curves have been plotted in terms of length of the antenna system expressed in wave lengths rather than in frequency, since in this way the curves are made perfectly general and apply to any antenna system. The relation between frequency $f$ and length expressed in wave lengths is

$$\text{Wave lengths} = \frac{(\text{actual length in meters})\ (\text{frequency in cycles})}{300{,}000{,}000} \tag{55}$$

Among the outstanding properties illustrated in Fig. 144 is the fact that the impedance curve repeats every half wave length, that maxima occur at even quarter wave lengths and minima at odd quarter wave lengths, and that the impedance is resistive whenever the length is exactly an integral multiple of a quarter wave length. Furthermore the reactance changes from capacitive to inductive at each quarter wave length point.[2]

The importance of the antenna impedance arises from the fact that the tuning reactance required to bring the antenna system into resonance is equal in magnitude but opposite in sign to the

[1] Methods of studying selective fading by the use of modulated waves are described by Ralph Bown, De Loss K. Martin, and Ralph K. Potter, Some Studies in Radio Broadcast Transmission, *Proc. I.R.E.*, vol. 14, p. 57, February, 1926; Ralph K. Potter, Transmission Characteristics of a Short-wave Telephone Circuit, *Proc. I.R.E.*, vol. 18, p. 581, April, 1930, and E. Bruce and A. C. Beck, Experiments with Directivity Steering for Fading Reduction, *Proc. I.R.E.*, vol. 23, p. 357, April, 1935.

[2] These impedance characteristics follow from the nature of the voltage and current relations existing on the antenna, and can be worked out from conventional transmission-line theory.

reactance of the antenna system. Likewise the antenna resistance determines the coupling that is required to match a transmission line to the antenna.

The reactance of an antenna system can be readily determined by tuning with an inductance or capacity as required until resonance is obtained with a loosely coupled oscillator operating at the desired frequency. The tuning reactance required to accomplish this is equal in magnitude and opposite in sign to the antenna reactance. Resonance between the antenna and the oscillator can be determined by tuning for maximum response in

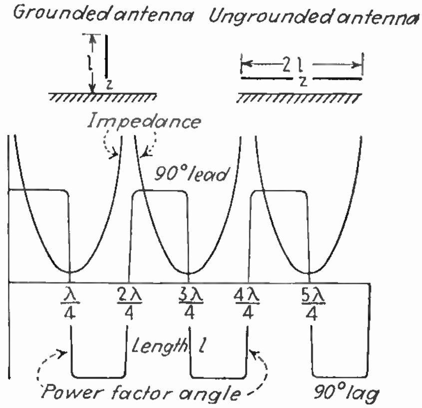



FIG. 144.—Antenna impedance and power-factor angle as a function of antenna length in wave lengths.

the antenna, or by noting the effect of the antenna tuning upon the oscillator plate current. Coupling between the antenna and the driving oscillator should be small. The resistance of an antenna system can be obtained by using the resistance-variation, reactance-variation, or resistance-neutralization methods of measuring radio-frequency resistance. The frequency-variation method cannot be used, however, because of errors that result as a consequence of the distributed character of the antenna constants.

The impedance of antennas and transmission lines can be measured by the substitution method if desired, by using a set-up of which Fig. 145 is a typical illustration. This shows an ordinary oscillator with which is associated a calibrated condenser $C$, a resistance box $R$ suitable for high-frequency service, and a wavemeter. A sensitive meter is also placed in the plate circuit and provided with a balancing arrangement whereby the normal plate current can be balanced out of the meter, which can therefore be made to detect small changes in plate current with considerable accuracy. When the impedance being measured is capacitive, the switch $S_1$ is first disconnected, while $S_2$ is closed. The oscillator is adjusted to the proper frequency and the d-c current balanced out of the plate-current instrument. The switch $S_2$ is now opened and $S_1$ is closed, thereby substituting the capacity $C$ and the resistance $R$ for the unknown impedance. The capacity $C$ is adjusted until the original frequency has been restored, after which the resistance $R$ is varied until the original

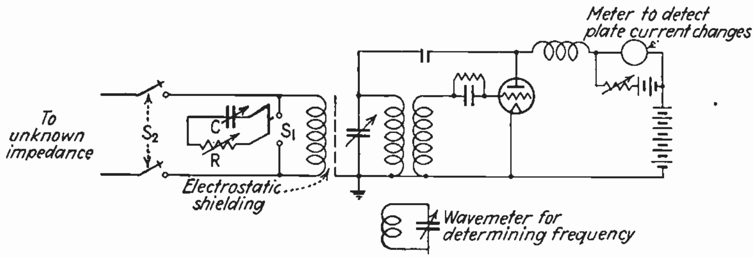



FIG. 145.—Typical example of a set-up for measuring antenna or transmission-line impedance by the substitution method.

losses are obtained as indicated by the original oscillator plate current. The unknown impedance is then obviously equivalent to $C$ and $R$ in series. If the impedance is inductive, one closes both $S_1$ and $S_2$, sets $R$ to zero, and adjusts for proper frequency as before. The unknown impedance is then disconnected by opening switch $S_2$, after which the capacity is readjusted to restore the original frequency, and the resistance $R$ is given a value that restores the original plate current. Then if $\Delta C$ is the change in capacity and $C$ the capacity after opening $S_2$,

$$\text{Unknown reactance} = \left(\frac{1}{\omega \Delta C}\right)$$

$$\text{Series resistance of unknown} = R\left(\frac{C}{\Delta C}\right)^2$$

The resistance of an antenna system is made up of dielectric losses, conductor resistance, eddy-current losses in neighboring objects, ground resistance, and radiation resistance. No satisfactory method has been devised for separating these quantities beyond taking advantage of the fact that for some antenna systems under idealized conditions it is possible to calculate the radiation resistance. Some effort has been made to separate the components by assuming each varies with frequency according to a different law, and then making measurements at several frequencies. This is open to the objection that the assumed laws of variation are at best only approximations because the voltage and current distribution in an antenna vary with the frequency.

*Resonant Frequency of Antennas.*—A resonant frequency of an antenna is defined as a frequency for which the antenna impedance is a pure resistance, and always occurs when the antenna

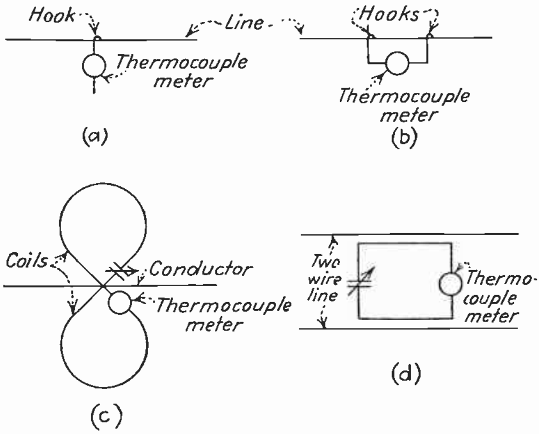



FIG. 146.—Devices for measuring the current distribution along transmission lines and radio antennas.

impedance is either a minimum or a maximum, as shown in Fig. 144. The resonant frequency can be determined by loosely coupling an oscillator to the antenna and varying the frequency of oscillations until resonance with the antenna is indicated either by maximum response of the antenna or by the reaction of the tuned antenna circuit upon the d-c plate current of the oscillator. The response of the antenna system can be observed either by placing a thermocouple meter in the antenna circuit or by coupling a sensitive thermocouple to the antenna conductor using a technique similar to that shown in Fig. 146.

In making resonant-frequency determinations, the coupling impedance inserted in series with the antenna under test should be a minimum, since any such impedance affects the resonant frequency obtained. For very accurate determinations, one can either excite the antenna by radiated waves produced by a transmitting antenna a few wave lengths away or determine resonant frequency for several values of added reactance and then extrapolate the resulting curve of frequency as a function of added reactance to zero reactance.

Every antenna has an infinite number of resonant frequencies, as is apparent from Fig. 144, but usually only one or two of these are of importance. Care must be taken in making measurements to be sure that the resonant frequency being measured is the one desired. This can be ascertained from a knowledge of the antenna dimensions or by determining the current distribution in the antenna at the frequency in question.

*Current Distribution.*—The current distribution along an antenna can be measured by the same technique employed to determine the current distribution in radio-frequency transmission lines, as discussed in Sec. 67. Current-distribution measurements are of particular importance in connection with complicated antenna systems and are commonly depended upon in tuning up antenna arrays to indicate whether or not the current nodes come at the right places.

*Effective Height.*—The effective height of an antenna can be defined as the ratio of the equivalent lumped induced voltage that can be thought of as acting in the antenna system divided by the field strength of the wave that induces this voltage. An alternative definition is to consider the effective height as the length of wire which, when carrying a uniform current $I$, will produce the same radiated field at a receiving point as is produced by the same current $I$ flowing at some reference point in the actual antenna. These two definitions lead to the same result and either can be used in making effective-height determinations.

In measuring effective height of transmitting antennas the usual procedure is to measure the field strength produced at a point a few wave lengths away. One then has from fundamental radiation theory

$$\text{Effective height in meters} = \frac{d\lambda\epsilon}{188.5I} \qquad (56)$$

where

$d$ = distance to antenna in meters

$\lambda$ = wave length of radiated wave, in meters

$\epsilon$ = observed field strength in volts per meter

$I$ = current in transmitting antenna in amperes.

The effective height of receiving antennas can be determined by measuring the voltage which a wave of known strength induces in the antenna, using the field-strength measuring technique described in Sec. 59. The waves employed can be produced by either a portable transmitter or a near-by radio station, and the equipment for measuring the wave strength and induced voltage can be quite simple as strong fields may be used. In making measurements it is necessary that the transmitting and receiving points be separated by at least several wave lengths to insure plane waves and to avoid errors from induction fields. Where it is more convenient to do so, the procedure may be modified by using the transmitting-antenna technique to measure the effective height of receiving antennas, and vice versa.

Since the current in a transmitting antenna is different at different parts of the antenna, and since likewise the equivalent lumped voltage that can be thought of as producing the same effect as the actual radio wave also depends upon the position in the antenna at which this voltage is assumed to act, it is apparent that the effective height of an antenna will depend upon the position to which the results are referred. The usual convention is to refer the effective height to a point of maximum current in an ungrounded antenna and to the grounding point in a grounded antenna. It is also to be remembered that the radiation from a transmitting antenna depends upon the direction, and likewise that the effective voltage induced in a receiving antenna depends upon the direction of arrival of the waves. The effective height is therefore different in different directions from the same antenna.

*Directional Characteristics of Antennas.*—The directional characteristics of an antenna are the same when the antenna is used to transmit as when it is used to receive. Hence one can determine the directional characteristics either by radiating power from the antenna and measuring the field-strength distribution that results, or by measuring the voltage induced in the antenna as a portable transmitter is moved about in different directions. The former arrangement is usually preferred as being the most convenient. In carrying out the measurements, care must be taken to keep the distance between transmitter and receiver at least one wave length to avoid errors from induction fields.

A complete directional characteristic requires the use of an airplane, and measurements taken from the ground have only limited significance since they give only the field-strength distribution in the horizontal plane of the vertically polarized component of the ground wave. This is of importance at broadcast and lower frequencies, but at higher frequencies the important thing is the space distribution of radiation at angles somewhat above the horizontal. Furthermore horizontally polarized antennas are commonly employed at high frequencies and these frequencies produce virtually no field strength at the surface of the ground.

*Antenna Power.*—The power delivered to an antenna can be determined from a knowledge of the antenna resistance and current. It is also possible to determine power with the three-ammeter shunt-impedance method described in Sec. 8, and also from measurements made upon the output power amplifier or oscillator with the antenna connected and disconnected as described in Sec. 55.

The harmonic power delivered to an antenna can be determined by using the technique described in Sec. 39.

*Model Antennas.*—The properties of an antenna can sometimes be investigated to advantage by the use of models, particularly when the antenna about which information is desired is physically very large. The idea is to make a small-scale model of the antenna which is operated at a frequency such that the dimensions of the model measured in wave lengths will be the same as the dimensions of the full-sized antenna when measured in wave lengths at the operating frequency. The model antenna will then have the same radiation resistance, directive characteristics, field distribution about the antenna, etc., as will be obtained with the full-sized antenna, and by suitable adjustment of resistances the efficiency can likewise be made identical. It is possible in this way to obtain accurate information concerning the effect of guy wires, towers, counterpoise design, etc., and to compare alternative designs.[1]

[1] See G. H. Brown and Ronold King, High-frequency Models in Antenna Investigations, *Proc. I.R.E.*, vol. 22, p. 457, April, 1934; H. P. Miller, Jr., The Insulation of a Guyed Mast, *Proc. I.R.E.*, vol. 15, p. 225, March, 1927. H. E. Gihring and G. H. Brown, General Considerations of Tower Antennas for Broadcast Use, *Proc. I.R.E.*, vol. 23, p. 311, April, 1935.

The fact that an antenna operated at one frequency is equivalent to a larger antenna operated at a lower frequency can be used to advantage in obtaining optimum proportions in tower radiators such as those used in broadcast transmitters. Thus the optimum height of a tower radiator is approximately $\lambda/2$ but varies somewhat with the tower cross-section, ground-wave attenuation, etc. The simplest way to insure obtaining the optimum height for any given tower is to make periodical measurements on tower performance as a function of frequency during the construction. For each height there will be an optimum frequency, and from the trend of several such determinations at increasing physical heights one can readily anticipate what the proper height for the operating frequency will be. The construction can thus be stopped at the proper point and all costly experimentation and mistakes avoided.[1]

**66. Reciprocal Relations between the Transmitting and Receiving Properties of Antennas.**—The properties of an antenna when used to abstract energy from a passing radio wave are similar in nearly all respects to the corresponding properties of the same antenna when acting as a radiator. Thus the directional characteristics, the current distribution, the effective height, and the impedance of the antenna are the same in transmission as in reception. The only difference is in the radiation resistance, which in the case of receiving antennas depends upon the inserted load impedance and tends to be higher than when radiating.[2] These reciprocal relations between the transmission and reception properties are extremely useful because they make it possible to deduce the merits of a receiving antenna from transmission tests, and vice versa. Thus when radio waves are used

[1] The author is indebted to Mr. K. G. Ormiston of radio station KNX for a description of this method of determining the optimum height of tower antennas. Station KNX erected in 1934 the first half-wave-type tower antenna having a wide base. A testing procedure like that described above showed that for this type of tower the optimum height was 0.49 wave length, and the construction was stopped at this point. All tower radiators in use at that time had a narrow base and were from 0.54 to 0.62 wave length high. If the wide base tower had been blindly built to such a height, the performance would have been distinctly less than optimum, and it would have been an expensive matter to reduce the tower height.

[2] See Raymond M. Wilmotte, Generalized Theory of Antennae, *Exp. Wireless and Wireless Eng.*, vol. 5, p. 119, March, 1928.

as a means of obtaining a bearing, the accuracy is the same when the ship which desires to determine its location observes the direction of the waves sent out from a radio station as it is when the operator at the radio station observes the direction of signals sent out from the ship. Other illustrations of the application of the reciprocity principle have already been discussed in connection with the determination of effective height and the directional characteristics of antennas.

These reciprocal relations between the transmitting and receiving properties of an antenna are incorporated in two reciprocal theorems. The first of these was discovered by Rayleigh and extended to include radio communication by John R. Carson. It is to the effect that if an e.m.f. $E$ inserted in antenna 1 causes a current $I$ to flow at a certain point in a second antenna 2, then the voltage $E$ applied at this point in the second antenna will produce a current $I$ at the point in antenna 1 where the voltage $E$ was originally applied. The second theorem is due to Sommerfeld and Pfrang and states that with two antennas $A$ and $B$ of arbitrary orientation, the average phase and intensity of the field at $B$ when $A$ is transmitting with a given average power are the same as the average phase and intensity of the field at $A$ when $B$ is transmitting with the same average power.[1] The Rayleigh-Carson theorem fails to be true only when the propagation of the radio waves is appreciably affected by an ionized medium in the presence of a magnetic field, and so holds for all conditions except short-wave transmission over long distances. The Sommerfeld-Pfrang theorem, on the other hand, in addition to having the same limitations as the Rayleigh-Carson theorem, also fails to hold when the antennas are near the earth.

**67. Measurements on Transmission Lines.**—The impedance offered by a transmission line can be measured by the substitution method, by using the same technique and equipment discussed in connection with Fig. 145 for the measurement of antenna impedance.

The current distribution along a transmission line indicates whether or not proper operating conditions are realized. Thus in a non-resonant line, the current should die away exponentially

[1] For a discussion of the Rayleigh-Carson and Sommerfeld-Pfrang theorems see John R. Carson, Reciprocal Theorems in Radio Communication, *Proc. I.R.E.*, vol. 17, p. 952, June, 1929.

from the sending end and the presence of any resonance indicates improper adjustments. Similarly the current distribution in resonant lines will indicate whether the resonance is of the character desired.

Current distribution can be measured by coupling a sensitive thermocouple instrument to the line. Various arrangements suitable for this purpose are illustrated in Fig. 146. In the device shown at Fig. 146*a* the stray capacity between the free wire hanging from the thermocouple instrument and the transmission line supplies the return path for the current, and the amount of coupling to the transmission line depends upon the length of this wire. The arrangement at Fig. 146*b* operates from the voltage drop along the transmission line. The devices shown at *c* and *d* couple to the magnetic field produced by the wire currents, and have the advantage of avoiding contact with the transmission line. The arrangements shown at *a*, *b*, and *c* give an indication of the current in the individual wire involved, and unbalances between the two sides of the transmission line can be detected by comparing the relative currents of the two wires. In contrast with this, the device shown in Fig. 146*d* is sensitive only to the balanced component of the current, *i.e.*, responds only to the current which flows out one wire and returns through the other wire, and does not respond at all to currents which may flow out through the two wires acting in parallel and return through the ground.[1]

The attenuation of a transmission line is usually small and therefore difficult to measure accurately unless the line is long. The usual way of determining the attenuation is to terminate the line so that it is in a non-resonant condition and then calculate the attenuation constant from the observed ratio of sending-end current $I_s$ to receiving-end current $I_r$ according to the formula

$$\text{Total line attenuation in decibels} = 20 \log_{10}\left(\frac{I_s}{I_r}\right) \qquad (57)$$

In making the determination of receiving- and sending-end currents it is desirable to make a number of measurements distributed over a distance of at least one-fourth wave length from each end, and then take average values for the terminal currents

[1] E. J. Sterba and C. B. Feldman, Transmission Lines for Short-wave Radio Systems, *Proc. I.R.E.*, vol. 20, p. 1163, July, 1932.

to eliminate errors from residual resonance. An alternative method of determining attenuation is either to short-circuit or to open-circuit the receiving end of the line and measure the sending-end impedance at a frequency for which the line is in exact resonance. The attenuation can then be calculated according to the formulas[1]

$$\text{Total attenuation in decibels} = 8.686\frac{Z_0}{Z_g} \tag{58a}$$

$$\text{Total attenuation in decibels} = 8.686\frac{Z_g}{Z_0} \tag{58b}$$

where $Z_0$ is the characteristic impedance of the line and $Z_g$ is the sending-end impedance at resonance. Equation (58*a*) applies when the line is an odd number of quarter wave lengths long and short-circuited at the receiver, or an even number of quarter wave lengths long and open at the receiver. Equation (58*b*) applies when the line is an even number of quarter wave lengths long and short-circuited at the receiver, or an odd number of quarter wave lengths long and open at the receiver.

The characteristic impedance of a line is usually determined by calculation from the physical dimensions involved. It can, however, be determined by measuring the impedance at the sending end of the line, first, when the receiver is open and, second, when the receiver is short-circuited. The characteristic impedance is then equal to $\sqrt{Z_s Z_p}$, where $Z_s$ and $Z_p$ are respectively the short-circuit and open-circuit impedances. In making an experimental determination of characteristic impedances it is desirable to make the measurements at a frequency for which the line is resonant, since then the impedances to be determined are pure resistances and can be evaluated with a minimum of experimental difficulty.

[1] See F. E. Terman, Resonant Lines in Radio Circuits, *Elec. Eng.*, vol. 53, p. 1046, July, 1934.

# CHAPTER XII

## LABORATORY OSCILLATORS

**68. Requirements for Laboratory Oscillators.**—Oscillators used to operate bridges and to make other laboratory measurements are normally required to have unusually good wave form and frequency stability. The most important factor contributing to good wave shape is limitation of the operating region to the linear part of the tube characteristic. High $Q$ resonant circuits, resonant circuits having a small $L/C$ ratio, and resonant circuits so arranged that the impedance to harmonics is as low as possible also help.

The frequency stability of an oscillator depends largely upon the extent to which changes in the electrode voltages of the tube can be prevented from affecting the relative phase of the voltages across the grid-cathode and plate-cathode sections of the tuned circuit. Any alteration of the relative phase of these voltages will cause the oscillator to operate slightly off the resonant frequency of the tuned circuit in order that the tuned circuit may introduce a compensating phase shift. Factors tending to prevent phase shifts with changes in electrode voltages are good wave form, small $L/C$ ratio, and, in the case of Hartley and feed-back oscillator circuits, the closest possible coupling between the grid and plate parts of the tuned-circuit inductance. Care must also be taken to prevent phase shifts between the tube and the tuned circuit by grid and blocking condensers that are too small, or shunt feed reactances with insufficient impedance. The effect upon the frequency of the phase shifts that cannot be avoided is inversely proportional to the $Q$ of the resonant circuit of the oscillator, since the higher the $Q$ the less the frequency change required to produce a given phase shift. When proper attention has been given to the foregoing factors, the principal cause of frequency variation is usually the temperature coefficient of the tuned circuit. This effect can be minimized by so locating the heat-producing elements of the oscillator (tubes, voltage dropping resistances, bias resistors, etc.) that the heating of the tuned circuit is a minimum.

The tuned-circuit details depend upon the frequency. Paper condensers are entirely satisfactory at audio frequencies, while mica fixed condensers and air-dielectric tuning condensers are preferred at radio frequencies. Coils for audio-frequency service may be wound on cores of audio-transformer steel provided with a small air gap, or may employ permalloy or iron-dust cores such as used in telephone loading coils. At frequencies from about 15,000 cycles up to about 500,000 cycles, the coils can be either bank- or universal-wound air-core coils, preferably using litz wire,

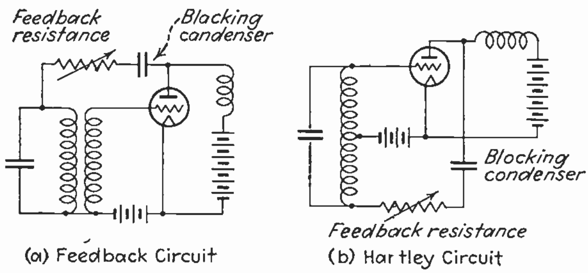



FIG. 147.—Typical oscillator circuits provided with resistance stabilization.

or they can be wound on dust cores especially designed for radio-frequency service (such as Polyiron and Ferrocart), which give good $Q$ and have the advantage of compactness.

A buffer tube between the oscillator and load is necessary (except with electron-coupled oscillators) in order to prevent the load from affecting the frequency of oscillations. This buffer tube can be a triode for audio-frequency oscillators but must be a screen-grid or pentode tube for radio-frequency oscillators.

**69. Resistance-stabilized Oscillators.**—The resistance-stabilized oscillator is widely used for the generation of audio and low radio frequencies. Typical circuit arrangements are shown in Fig. 147 and are seen to be conventional oscillator circuits with the addition of a "feed-back" resistance located between the plate of the oscillator tube and the tuned circuit. This feed-back resistance must be high compared with the plate resistance of the tube and has two primary functions. First, it makes the resistance which the tuned circuit sees when looking toward the plate substantially independent of the electrode voltages; and, second, it provides a means for limiting the amplitude of oscillations to the straight-line part of the tube characteristic.

The tube in a resistance-stabilized oscillator is adjusted to operate as a Class A amplifier, and the feed-back resistance is made so high that oscillations are just barely able to start. Under these conditions, oscillations start with minute amplitude and build up until there is grid current, which introduces additional losses that increase rapidly with further increase in amplitude. If the feed-back resistance is so high that oscillations are barely able to exist with no grid loss, an equilibrium will be reached at an amplitude which drives the grid only a few volts positive. It will be noted that a fixed grid bias such as obtained from a biasing resistance is necessary, and that the grid-leak bias arrangement commonly used with power oscillators is not permissible.

The wave form is determined by the linearity of the tube's dynamic characteristic over the range of voltage which the oscillations apply to the grid. It is apparent that for good wave form the tube when considered as an amplifier must be so adjusted that it will amplify without distortion an alternating-current voltage on the grid having a crest value slightly greater than the grid bias. This means that the oscillator tube should be operated at a grid bias that is slightly less than the bias that would be used for Class A amplifier operation at the same plate voltage.

Best results are obtained when attention is paid to certain circuit details. The circuit proportions should be such that the feed-back resistance required is at least twice, and preferably over five times, the plate resistance. The blocking condenser in series with the feed-back resistance must have a low reactance compared with this resistance in order to avoid phase shifts, while the shunt-feed choke should have a reactance that is high compared with the plate resistance of the tube for the same reason. The frequency stability is also helped greatly by making the coupling between plate and grid coils as close as possible. Two possible methods of connecting a buffer tube are shown in Fig. 148. The arrangement at Fig. 148*a* is usually preferred because it gives the best wave form, although the circuit of Fig. 148*b* has the advantage of developing greater output voltage.

The most satisfactory tubes for resistance-stabilized oscillators are those having amplification factors in the range of 4.5 to 8, together with the highest possible mutual conductance. With such tubes, the grid and plate coils should have approximately unity turn ratio.

*Design of Resistance-stabilized Oscillators.*—When the characteristics of the resonant circuit are known, it is possible to lay out a resistance-stabilized oscillator on paper and predict accurately the amplitude of oscillations and the circuit conditions required for proper operation. For example, assume that it is desired to set up an oscillator employing a tuned circuit that develops a parallel-resonant impedance between plate and filament taps of 50,000 ohms. Assume further that the ratio

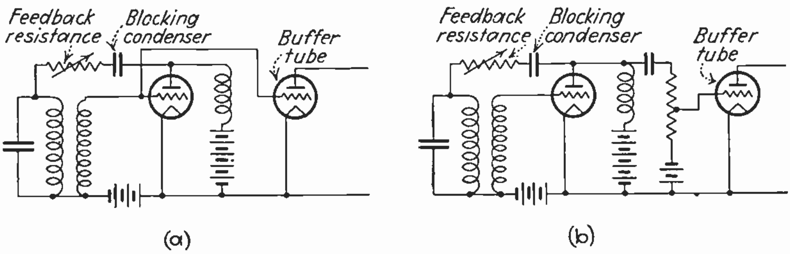



FIG. 148.—Methods of coupling a buffer tube to a resistance-stabilized oscillator.

between plate and grid coils is 1 to 1, that a Type 89 tube operated as a Class A triode amplifier is to be employed, and that an amplitude of oscillations of 10 volts crest is desired. The first step is to select a grid bias that will be 2 to 3 volts less than the crest amplitude, so that 7.5 volts bias will be satisfactory. The plate voltage is now chosen so that the operating region is located on a straight-line part of the tube characteristic. This calls for the highest plate voltage that will not give excessive plate current with the grid bias, which for this case is about 110 volts.

The feed-back resistance that will just barely enable oscillations to start has a value such that, when 1 volt is applied to the grid of the tube, exactly 1 volt will be developed across the tuned circuit by amplifier operation. If the impedance of the shunt-feed choke is very high compared with the plate resistance of the tube, the feed-back resistance at which oscillations will just start is given by the formula

$$\text{Starting feed-back resistance} = R_L(\mu - 1) - R_p \qquad (59)$$

where $R_p$ is the plate resistance of the tube, $R_L$ the load resistance offered by the tuned circuit, and $\mu$ the amplification factor of the tube. In the case at hand $R_L = 50{,}000$, while reference to a tube chart shows $\mu = 4.7$ and $R_p = 3000$. The critical feed-back resistance hence works out to be 182,000 ohms, and the value actually employed should be 5 to 15 per cent less, or roughly 165,000 ohms. In practice the resistance is usually adjusted experimentally by setting it at a value about 10 per cent less than that at which oscillations start, but calculations such as have been

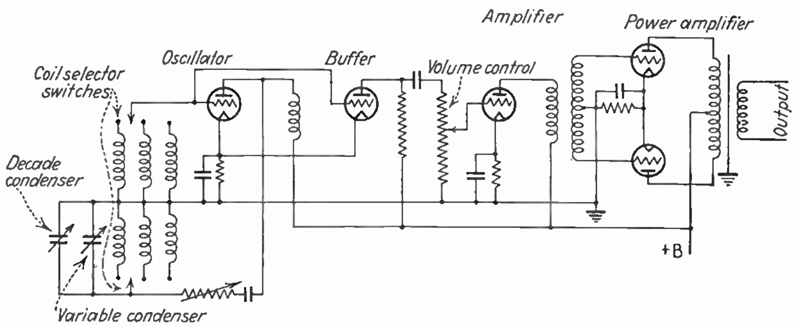



FIG. 149.—Circuit diagram of typical resistance-stabilized oscillator for covering the audio-frequency range.

outlined are of considerable aid in establishing the limiting values that will be needed.

From Eq. (59) it will be noted that the feed-back resistance that is required will be nearly proportional to the tuned-circuit resistance $R_L$, and this fixes limits to the allowable $L/C$ ratio in the tuned circuit since $R_L$ is proportional to $\sqrt{L/C}$ when the circuit $Q$ is constant. It is undesirable to use feed-back resistances higher than about 500,000 ohms at audio frequencies, and higher than 50,000 to 100,000 ohms at the lower radio frequencies. At the same time, the $L/C$ ratio must not be too low, since it is desirable that the feed-back resistance be at least twice, and preferably over five times, the plate resistance of the tube. The most suitable values of tuned-circuit resistance are in the range 10,000 to 50,000 ohms.

A complete circuit diagram of a resistance-stabilized oscillator for generating audio and carrier frequencies to be used in laboratory measurements is shown in Fig. 149. Continuous variation of frequency is obtainable by using a continuously variable tuning condenser consisting of a decade condenser supplemented by a variable air condenser to interpolate between the smallest steps, together with provision for switching various coils in and out of the circuit. The feed-back resistance can be a tapped or adjustable commercial wire-wound resistance arranged with a tap switch so that the feed-back resistance can be varied in increments of about 10 per cent. The proper setting of the feed-back resistance will depend upon the frequency and can be given on the frequency-calibration chart.

The resistance-stabilized oscillator is the most satisfactory type of tuned-circuit oscillator available for generating audio frequencies in the laboratory. The amplitude of oscillations is constant over the entire frequency band (assuming the feed-back resistance is readjusted as necessary), the wave form is practically perfect except for distortion that may be introduced by the output amplifier, and the frequency is practically independent of tube voltages and tube replacements. Resistance-stabilized oscillators are also simple to build and easy to adjust. They are used primarily for audio and carrier frequencies and can be employed up to 100 to 200 kc. At frequencies higher than this, stray capacities tend to by-pass the feed-back resistance and thereby nullify the advantages of the circuit.

**70. Dynatron Oscillators.**[1]—In the dynatron oscillator, the negative plate-filament resistance of a dynatron tube is connected in parallel with the resonant circuit, as shown in Fig. 150. A dynatron tube has a characteristic such as shown in Fig. 51, and over an appreciable range of plate voltages exhibits a substantially constant negative plate-cathode resistance of a magnitude that depends upon the total space current.[2]

[1] For further information see Janusz Groszkowski, Oscillators with Automatic Control of the Threshold of Regeneration, *Proc. I.R.E.*, vol. 22, p. 145, February, 1934.

[2] The negative-resistance characteristic is a result of secondary electron emission at the plate and is obtained when the grid next to the plate is more positive than the plate, and provided the plate voltage and surface conditions of the plate electrode are such that appreciable secondary emission takes place. For further information see "Radio Engineering," Chap. IX.

Oscillations start when the absolute magnitude of the negative resistance[1] is less than the parallel-resonant impedance of the tuned circuit, and they build up in amplitude until the straight-line part of the negative-resistance region is exceeded. The operating range then extends into a region where the negative resistance is higher, thus increasing the "average" or effective resistance of the tube until the effective negative resistance is equal to the parallel-resonant impedance of the tuned circuit, at which amplitude an equilibrium is established. The dynatron oscillator performs best when the operating point is in the middle of the negative resistance region and the amplitude of oscillations is such as to confine the operating range as nearly as possible to the straight-line part of the tube characteristic. Under these conditions the wave form is practically sinusoidal, and the frequency stability is likewise highest. In a simple dynatron oscillator such as is illustrated in Fig. 150, the proper adjustment is therefore such that the grid bias is the most negative value that will permit oscillations to start, since under these conditions an equilibrium will be obtained before the operating region gets appreciably beyond the straight-line part of the tube characteristic.

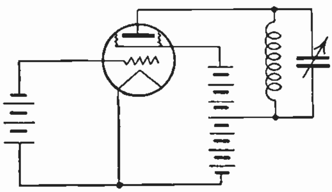


FIG. 150.—Circuit of simple dynatron oscillator.

The advantage of restricting the amplitude of the oscillations to the straight-line part of the tube characteristic makes it desirable to employ an automatic amplitude- or volume-control arrangement such as shown in Fig. 151. Then, as the amplitude of oscillations increases, the negative bias upon the grid of the dynatron tube increases, thus increasing the negative resistance and reducing the tendency to oscillate.

With proper circuit proportions it is possible to obtain equilibrium before the oscillations have reached the limit of the straight-line part of the tube characteristic. This insures the best possible wave form, high frequency stability, and also makes the oscillator self-adjusting in that, as the resonant frequency of the tuned circuit is varied, thus changing the parallel imped-

[1] In the discussion that follows all references to the magnitude of the negative resistance mean absolute magnitudes. That is, a larger negative resistance means a larger negative number.

ance, the oscillator automatically readjusts itself to maintain proper operating conditions. Furthermore, the oscillations have nearly the same amplitude for wide variations in the impedance of the tuned circuit.

The automatic amplitude control may be a simple automatic volume-control system using a diode as shown in Fig. 151*a* or an arrangement such as illustrated in Fig. 151*b* which gives amplified volume control. Delay action[1] can be introduced in Fig. 151*a* by biasing the anode of the diode negative, and in Fig. 151*b* by making the bias on the automatic volume-control tube

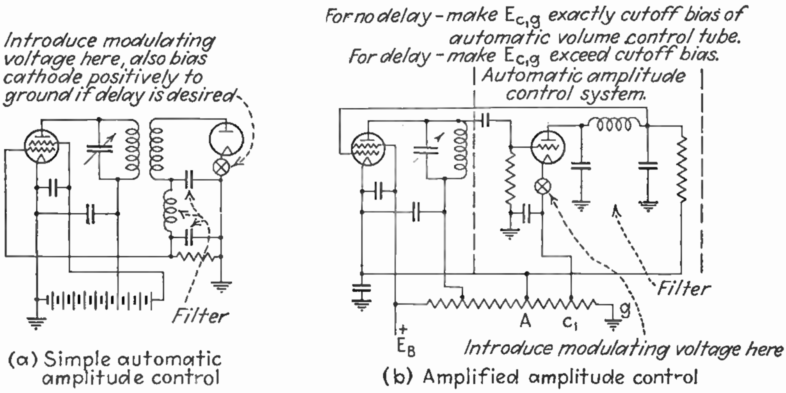



FIG. 151.—Dynatron oscillators provided with automatic amplitude control.

exceed cut-off. When the volume control possesses both amplification and delay, the amplitude of oscillations will be very nearly constant at all times, and the advantages of amplitude control are realized to the greatest extent.

A dynatron oscillator provided with automatic volume control has practically perfect wave form and remarkably high frequency stability with variations in electrode voltages. Such oscillators will operate at any frequency up to about 30 mc, with the upper limit being fixed in part by the fact that it is difficult to obtain a negative resistance low enough to make ultra-high frequency circuits oscillate, and in part by the fact that above 30 mc the time of flight of the electrons cannot be ignored. The only

[1] An automatic volume control is said to possess "delay" when arranged so that it does not begin to operate until the volume has reached some predetermined level. Amplified automatic volume control is obtained when the volume-control effect is amplified in some manner.

disadvantage of the dynatron oscillator is that the negative-resistance characteristics of different tubes vary greatly and also that the characteristic of an individual tube will often change with age. The dynatron oscillator provided with automatic volume control represents what is probably the best type of oscillator for laboratory use at radio frequencies. It can also be used at audio frequencies, but the resistance-stabilized oscillator gives as good a performance and is much more stable with respect to tube characteristics.

**71. Conventional Oscillator with Automatic Amplitude Control.**[1]—Any conventional oscillator circuit can be made to generate oscillations having good wave form and high frequency

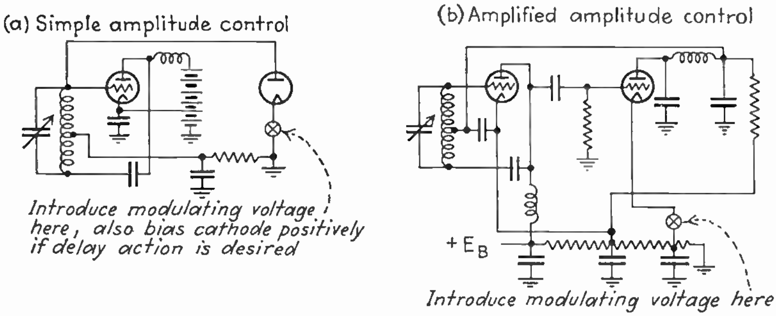



FIG. 152.—Conventional Hartley oscillator with automatic amplitude control.

stability by operating the tube as a Class A amplifier, and using an automatic amplitude-control arrangement to limit the oscillations to the straight-line part of the tube characteristic. The volume-control circuits given in Fig. 151 can be adapted to this type of oscillator as shown in Fig. 152. When the oscillations start, the grid bias is either zero or very small so that the plate resistance is a minimum; but as the oscillations build up, the negative bias on the tube becomes greater, thus shifting the operating point toward regions of higher plate resistance where the tendency to oscillate is less. With proper circuit proportions, equilibrium conditions can be obtained with an amplitude of oscillations that is small, and at an operating point corresponding to Class A operation of the tube. The result is excellent wave

[1] These oscillators are discussed in greater detail by L. B. Arguimbau, An Oscillator Having a Linear Operating Characteristic, *Proc. I.R.E.*, vol. 21, p. 14, January, 1933.

shape, good frequency stability, and substantially constant amplitude of oscillations as the frequency of oscillations is varied. These advantages are realized to the greatest extent when the automatic volume control is of the type having both delay and amplification. Any of the ordinary oscillator tubes can be employed although there is some advantage in using variable-mu pentode or screen-grid tubes.

The conventional oscillator with automatic amplitude control is particularly suitable for use at radio frequencies, and is a direct competitor of the dynatron oscillator in this application. Data on performance are not available, but it appears that both frequency stability and wave form are slightly, although not

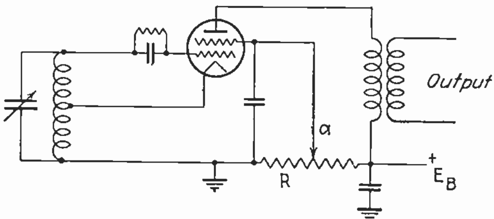



FIG. 153.—Typical electron-coupled oscillator circuit with grounded screen grid. The frequency can be made independent of plate voltage by connecting the screen-grid tap *a* to proper point on the voltage-dividing resistance *R*.

seriously, inferior to those obtained with the dynatron oscillator. At the same time the tube characteristics are stable, and the highest frequency at which oscillations can be obtained is greater.

**72. Electron-coupled Oscillators.**[1]—The circuit of a typical electron-coupled oscillator is shown in Fig. 153. This is a conventional triode oscillator circuit in which the screen grid serves as the oscillator anode while the plate functions as an output electrode. The load impedance in the plate circuit is prevented from reacting back on the oscillator by grounding the screen grid to alternating potentials and allowing the cathode to float at a potential above ground. In this way the screen serves as a grounded shield between the oscillator and output portions of the tube, while allowing most of the electron stream to reach the plate. It will be observed that when the plate electrode is completely shielded electrostatically the only coupling between the oscillator part of the tube and the output is through the

[1] See J. B. Dow, A Recent Development in Vacuum Tube Oscillator Circuits, *Proc. I.R.E.*, vol. 19, p. 2095, December, 1931.

electron stream. Hence when the plate voltage is high enough so that the plate current is not affected by plate voltage, then the oscillator portion of the tube delivers energy to the load in the plate circuit, but the plate circuit cannot react back on the oscillations. The only source of reaction in practical cases is through unshielded electrostatic couplings in the stem and base of the tube, and in practice these are very small.

The variation of frequency with plate and screen voltages in the electron-coupled oscillator depends primarily upon the ratio $E_{sg}/E_p$ *and there is always some value of this ratio for which the frequency is independent of the voltage.* High frequency stability hence can be insured by obtaining the screen voltage from a voltage divider as shown in Fig. 153, and selecting the location of the tap by trial so that the frequency does not change when the potential applied to the voltage divider is varied. The proper location of this tap may change slightly as the resonant frequency of the tuned circuit is altered, but a particular adjustment is quite satisfactory over a considerable frequency range.

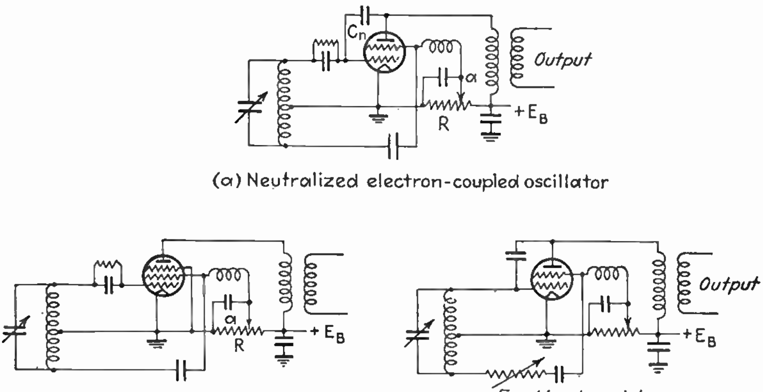
The electron-coupled oscillator normally operates with a grid bias considerably greater than cut-off, so that the plate current is in the form of pulses which are rich in harmonics. For this reason the electron-coupled oscillator is particularly satisfactory for use in heterodyne wave meters, since the frequency stability is extremely high, numerous harmonics are present in the output, and no buffer tube is required. Such electron-coupled oscillators also operate at high plate efficiency, although this is not particularly important in most laboratory oscillators.

The electron-coupled principle can be employed in a number of ways in addition to the arrangement illustrated in Fig. 153. Several of the variations most useful in laboratory work are shown in Fig. 154, and the pentagrid-converter circuits sometimes used in broadcast receivers use electron-coupled oscillators.[1] The oscillator of Fig. 154*a* differs from Fig. 153 only in that the cathode is grounded, thus permitting the use of filament-type tubes. Reaction of the output circuit on the oscillator portion of the tube is prevented by the neutralizing condenser $C_n$. In

[1] Most pentagrid-converter arrangements do not completely isolate the oscillator from the output circuits. See Paul W. Klipsch, Suppression of Interlocking in First Detector Circuits, *Proc. I.R.E.*, vol. 22, p. 699, June, 1934.

Fig. 154*b* a pentode has been substituted for the screen-grid tube. This arrangement does not require a neutralizing condenser because of the shielding action of the suppressor grid, but has the disadvantage that normally no position for the screen-grid tap can be found on the voltage divider that will make the frequency independent of anode voltage.

The oscillator of Fig. 154*c* is a resistance-stabilized electron-coupled oscillator. The oscillator portion is designed exactly as in the case of an ordinary resistance-stabilized oscillator, by




(a) Neutralized electron-coupled oscillator

(b) Pentode electron-coupled oscillator

(c) Resistance-stabilized electron-coupled oscillator

Fig. 154.—Various types of oscillators employing the electron-coupled principle.

using a fixed bias such that the operation is on the linear part of the tube characteristic, and employing the highest feed-back resistance that will permit stable operation. The output is then obtained from the anode circuit, and can be prevented from affecting the oscillator frequency by neutralization as shown. The resistance-stabilized electron-coupled oscillator combines the good wave form of the resistance-stabilized oscillator with the isolated output obtained by electron coupling. The frequency stability is obviously all that could be desired.

**73. Impedance-stabilized Oscillator.**[1]—It was explained in Sec. 68 that oscillators normally operate at a frequency differing slightly from the resonant frequency of the tuned circuit in order to introduce a compensating phase shift, and that this was

[1] See F. B. Llewellyn, Constant Frequency Oscillators, *Proc. I.R.E.*, vol. 19, p. 2063, December, 1931.

the principal reason for variation of frequency with supply voltage. This effect can be minimized by inserting a suitable reactance in series with the plate or grid electrodes of the tube to produce the required phase shift, thus permitting the oscillations to take place at exactly the resonant frequency of the tuned circuit. The amount of reactance which is required to do this can be shown to be substantially independent of the electrode voltages, so that if a suitable compensating reactance is used the oscillator always operates at exactly the resonant frequency of the tuned circuit and need not shift its frequency as the electrode voltages of the tube change.

If the resistance of the tuned circuit is zero, it is possible to make this compensation perfect, while with practical values of circuit resistance the compensation, although not complete, is very nearly so. The amount and nature of the reactance which must be employed depend upon the type and constants of the oscillator circuit. Formulas for calculating the compensating impedance for typical cases are given in Fig. 155 on the assumption that the tuned circuit has zero losses. When losses are present, the values giving the best stability will differ slightly from those computed, and must be determined by trial. It is always found that the best frequency stability is obtained when the circuit proportions are such that the required compensating reactance has a small size. An inspection of the equations given in Fig. 155 shows that this is when the ratio $L/C$ is small, and in the case of the Hartley and feed-back circuits also when the coupling between the grid and plate coils is as close as possible.

The impedance-stabilized oscillator is excellent for operation at a particular frequency but has the disadvantage that the stabilizing reactance required depends on the frequency and so must be changed as the tuning of the resonant circuit is varied. Since this cannot be conveniently done at radio frequencies, the impedance-stabilized oscillator is, in general, less satisfactory for laboratory use than is the electron-coupled oscillator.

**74. Modulation of Laboratory Oscillators.**—When laboratory oscillators operating at radio frequencies are to be modulated, the principal requirements are linearity of modulation, freedom from frequency modulation, and ability to follow the highest modulating frequencies to be encountered. While a modulated-amplifier arrangement can always be employed, it is much

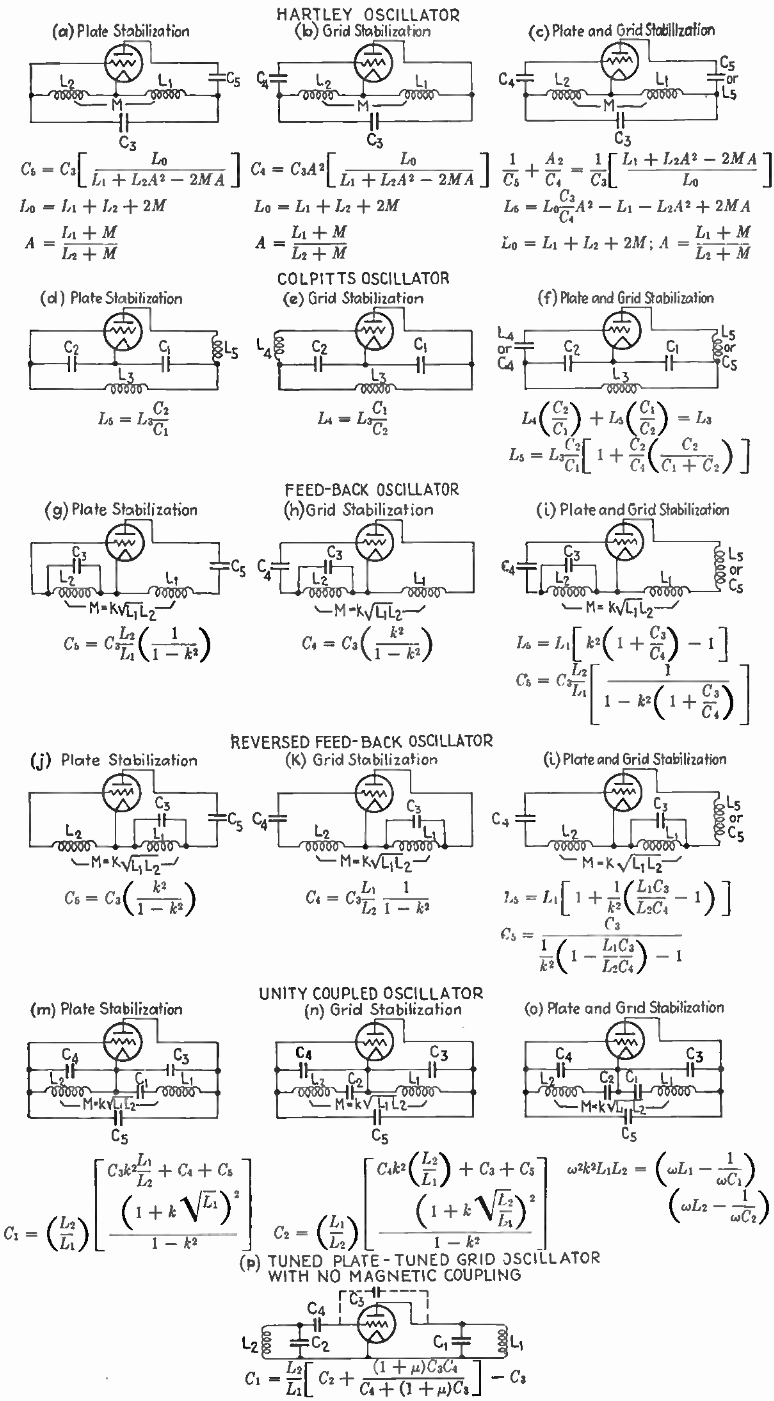



FIG. 155.—Circuits and circuit proportions that will make the generated frequency independent of the tube constants, provided the $Q$ of the resonant circuit is very high (theoretically infinite).

simpler, and usually permissible, to modulate laboratory oscillators directly, as described below.

The dynatron oscillator provided with automatic volume control, and likewise the conventional oscillator with automatic amplitude control, can be most satisfactorily modulated by introducing the modulating voltage in series with the bias developed by the automatic volume-control arrangement. This is indicated in Figs. 151 and 152 and the resulting modulation is linear if the automatic volume-control rectifier is linear, irrespective of whether or not delay action is present, provided that the automatic volume-control potentials can faithfully follow the modulating frequency.

The operation of this system of modulation in the absence of delay can be understood from the following example: Assume that the automatic volume-control arrangement develops one volt bias for each alternating volt developed across the resonant circuit, and assume further that equilibrium is obtained when the bias on the oscillator tube is $-10$ volts. When the modulating voltage is zero, the amplitude of the oscillations is then 10 volts. However, if a modulating potential of $-5$ is inserted in series in the circuit, the amplitude of oscillations will be such that the automatic volume control contributes $-5$ volts bias, since the equilibrium condition requires a total bias from all sources of $-10$. The amplitude of oscillations corresponding to this condition is 5 volts, provided the automatic volume-control rectifier is linear. Similarly when the modulating voltage is $+5$ volts, the automatic volume control must supply $-15$ volts bias in order that the equilibrium bias of $-10$ be present, and this correspond to oscillations of 15 volts amplitude. A little reflection shows that the amplitude of oscillations will follow the modulating voltage, provided the automatic volume-control bias is able to vary at the modulating frequency. With proper adjustments the frequency modulation will be negligible, and the modulation will be substantially distortionless. When delay action is present, the situation is very much the same except that the magnitude of modulating voltage required for complete modulation may be affected by the presence of the delay potential.

An impedance-stabilized oscillator can be plate modulated in the usual manner, as illustrated in Fig. 156*a*. The frequency modulation will be small since the frequency which is generated is substantially independent of plate voltage, provided the correct compensating impedance is employed. The degree of modulation obtainable depends upon the power output which the modulating tube can deliver. Complete modulation requires over three times as much tube capacity in the modulator as in the oscillator, and proper attention to the coupling between

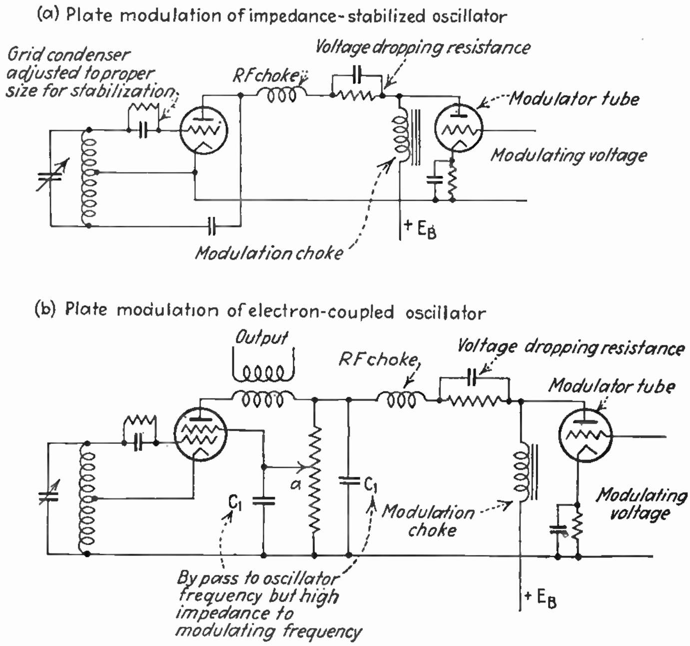



FIG. 156.—Plate modulation of electron-coupled and impedance-stabilized oscillators.

modulator and oscillator. In particular, when direct coupling is employed, the oscillator voltage must be reduced by a thoroughly by-passed voltage-dropping resistance consuming about 30 per cent of the plate potential (see Fig. 156).[1]

The resistance-stabilized oscillator cannot be modulated satisfactorily because the conditions under which it normally operates prevent the amplitude of oscillations from being able to follow any but the very lowest modulation frequencies.

[1] For a detailed discussion of the problems involved in plate modulation, see "Radio Engineering," pp. 361–366.

Plate modulation applied as shown in Fig. 156*b* represents the best method of modulating an electron-coupled oscillator. When the condensers marked $C_1$ are a by-pass to the oscillator but not the modulating frequency, this arrangement preserves the correct ratio between plate and screen voltages throughout the modulation cycle, thus avoiding the possibility of frequency modulation.

When the tuned circuit $Q$ is low enough to permit the amplitude of oscillations to follow the modulation frequency without discriminating against side bands, and when the time constant of the associated circuits such as automatic volume-control system, grid-leak—grid-condenser combinations, are such that it is possible for these to follow the modulating frequency, the

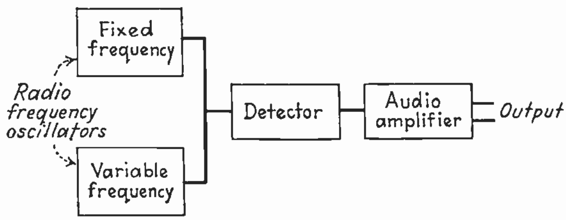



FIG. 157.—Schematic diagram of beat-frequency oscillator.

actual degree of modulation obtained and the linearity of modulation can be readily determined by replacing the alternating modulating voltage by a direct-current potential. Thus when plate modulation is employed, one can measure the amplitude of oscillations as a function of d-c plate voltage and from the resulting curve determine what happens as an alternating modulating voltage varies the plate potential about the operating point. Similarly one can plot a curve of amplitude of oscillations as a function of an added grid bias in oscillators which employ automatic amplitude control to limit their amplitude, and from the result determine what happens when the added bias is a modulating voltage.

**75. Beat-frequency Oscillators.**—In the beat-frequency oscillator, voltages obtained from two oscillators operating at slightly different radio frequencies are combined and applied to a detector, thus producing a difference-frequency current that constitutes the output. This is shown schematically in Fig. 157. The usefulness of the beat-frequency oscillator arises from the fact that a small percentage variation in the frequency of one oscillator, such as obtained by a single turn of a dial controlling a variable condenser, will vary the "beat" or difference frequency continuously through the entire audio range from a few cycles per second to 15,000 cycles or more.

The principal factors to consider in the design of a beat-frequency oscillator are the frequency stability of the output, the avoidance of coupling between the radio-frequency oscillators, and the arrangement of circuits so that spurious beat notes or "whistles" are not produced. The frequency stability of beat-frequency oscillators tends to be poor because a small percentage change in the frequency of one oscillator will produce a large percentage change in the comparatively small difference frequency. It is therefore necessary to employ radio-frequency oscillator circuits that have high inherent frequency stability. It is also desirable to make the two oscillators as nearly identical as possible so that changes in supply voltages, temperature variations, etc., will affect both alike and so have a minimum influence upon the difference frequency. Furthermore, the radio frequency at which the oscillators are to work must be neither too high nor too low. If high, then a small percentage frequency instability will produce a correspondingly large percentage change in the beat frequency. Whereas if a low radio frequency is selected, then when the beat frequency is large the oscillators will be operating at frequencies that differ by a considerable percentage and so will not act the same with respect to changes in temperature and electrode voltages. With a low frequency it is also impossible to make $L/C$ small (as required for best frequency stability) and still obtain the necessary tuning range with variable condensers commercially available. Experience indicates that a frequency of approximately 100 kc is best.

In order that the calibration curve of difference frequency as a function of variable condenser setting may be as accurate as possible, arrangements are always provided for setting the frequency of the fixed oscillator to make some selected point on the calibration correct. This is sometimes accomplished by use of a reed which is resonant at some known frequency. The procedure is to set the dial at the point on the calibration curve corresponding to this frequency, start the reed vibrating with an electromagnet energized by the oscillator output, and then adjust the fixed oscillator for maximum reed amplitude. Instead of the reed, one can use the 60-cycle power frequency as a standard by setting the dial to the 60-cycle point, applying several volts of 60 cycles to the detector plate circuit, and adjusting the fixed oscillator until zero beat is indicated by a milliammeter in the detector plate circuit. When the radio-frequency oscillators are so well shielded that they will not synchronize until the difference frequency is less than one cycle, the calibrating can be accomplished by turning the dial to zero frequency and adjusting the fixed oscillator so that a meter in the detector plate circuit indicates zero beat. Immediately after an adjustment is made in one of these ways, the frequency calibration is very accurate, but with time the frequencies of the two oscillators may drift by unequal amounts. This makes it necessary to check the adjustment at regular intervals, particularly when the oscillators are just warming up.

The two radio-frequency oscillators must always be thoroughly shielded from each other and arranged so that a minimum of coupling is introduced when their outputs are combined at the detector. This is necessary because two oscillators coupled together will synchronize if the difference in frequency is not too great, and such action will make it impossible to produce low-frequency beats. If care is used in the shielding and coupling, it is readily possible to prevent synchronization down to about one cycle per second, which is far below the lowest audible frequency.

The output wave obtained from the beat-frequency oscillator should be a sine wave free of spurious beat notes. These spurious beat notes or whistles often appear when the output frequency is high, and are the result of high-order harmonics of the radio-frequency oscillators (such as the tenth harmonic of one oscillator and the eleventh harmonic of the other) heterodyning together. The effect is to produce a whistle that varies from the upper limit of audibility down to zero, and then to beyond audibility while the normal output frequency is varied through only a few hundred cycles. The production of whistles in the detector is avoided by making the radio-frequency voltages applied to the detector as nearly sinusoidal as possible, and taking especial care to make one of the frequencies (usually the output of the fixed oscillator) absolutely sinusoidal by the use of suitable filter circuits. The audio-frequency amplifier following the detector can be prevented from producing whistles by keeping out radio-frequency voltages with a good low-pass filter in the detector output, and by operating the detector so that its output

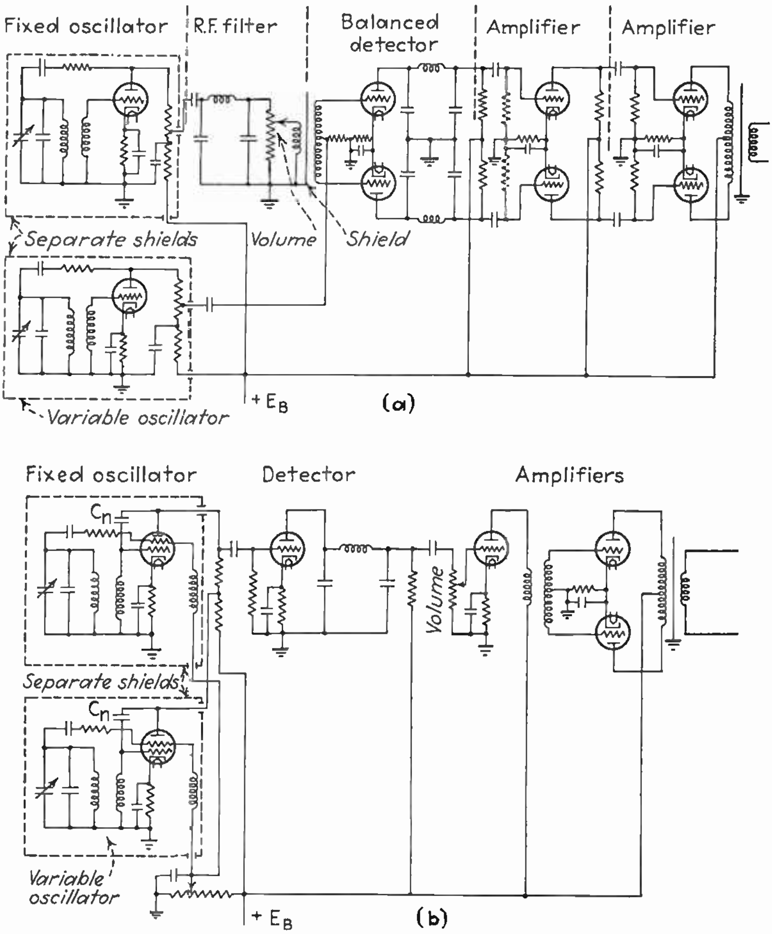



FIG. 158.—Circuit diagrams of typical beat-frequency oscillators.

contains as few harmonics as possible. The detector should thus be operated as a full-wave square-law rectifier, which means input voltages of only a few volts and a bias that is less than cut-off by an amount equal to the sum of the crest values of the two input voltages.

The best radio-frequency oscillators for use in beat-frequency oscillators are naturally those having high frequency stability with changes in electrode voltages and possessing good wave form. The ideal arrangement appears to be a resistance-stabilized electron-coupled oscillator such as illustrated in Fig. 154*c*, since this combines unusual frequency stability, good wave form, and isolated output in a single tube. Ordinarily resistance-stabilized oscillators are also very satisfactory, and at least one commercial beat-frequency oscillator utilizes dynatron oscillators.

Typical circuit arrangements for beat-frequency oscillators are shown in Fig. 158. In the arrangement of Fig. 158*a* the two radio-frequency oscillators are of the resistance-stabilized type. Coupling between them is prevented by the shielding and by-passing indicated, by obtaining the outputs from the plate circuits rather than from the resonant circuit, and by using the balanced-detector arrangement. Spurious beat notes are avoided by filtering the fixed frequency input to the detector, by suitable choice of detector grid-bias and input voltages, and by a filter in the detector output. The circuit shown in Fig. 158*b* employs resistance-stabilized electron-coupled oscillators. The simple arrangement shown for applying the radio-frequency voltages to the detector is permissible, since with proper neutralization there is no coupling between the plate circuits and the oscillator tuned circuit. A single detector-tube followed by suitable audio-frequency amplification ending up in a push-pull power amplifier is indicated. These two figures merely represent typical circuit arrangements that have been found satisfactory, and innumerable variations are possible, provided they carry out the general requirements outlined above for satisfactory operation.[1]

In the construction of beat-frequency oscillators, certain details must be given attention. The heat generated by tubes, resistors, etc., tends to warm the two radio-frequency oscillators at different

[1] A particularly satisfactory variation in the method of combining the two radio-frequency oscillator outputs is to use the Wunderlich tube as a bias detector, applying the voltages from the two oscillators separately to the two coplanar grids. The outputs are then combined in the plate circuits to give normal detector action, but since the output of each oscillator is applied to its own grid, there is very little coupling introduced between oscillators.

rates, resulting in slow frequency drifts that may take many hours to become stabilized. This difficulty can be avoided by physically arranging the tubes and other heat-producing elements so that the tuned circuits of the radio-frequency oscillators receive as little heating as possible. It is helpful to line the cans shielding the tuned circuits with balsa wood. Beat-frequency oscillators are capable of generating very low frequencies, and this often introduces trouble in the audio-frequency amplifier output transformers, since at extremely low frequencies, *i.e.*, 10 to 30 cycles, the cores of most output transformers will saturate, thereby introducing distortion. The frequency range obtainable is determined by the ratio of variable to fixed capacity used in tuning the variable-frequency oscillator. The variable condenser can be a three- or four-gang broadcast condenser with all sections in parallel. If desired, the low-frequency range may be spread out over the entire dial by cutting down one gang of the condenser so that its maximum capacity is only about 10 per cent of the total capacity, and then arranging a switch to cut out the remaining gangs when a restricted range is desired.

Beat-frequency oscillators are widely used in audio-frequency measuring work because they give a continuous sweep through the entire audio-frequency range with one turn of a dial. The wave form is practically sinusoidal when the oscillator is properly constructed, and the frequency stability is entirely satisfactory for most audio work, provided an initial adjustment is made to check one point on the calibration. Finally the oscillator delivers substantially constant output through the entire frequency range to the extent that the output amplifier introduces no frequency discrimination.

## CHAPTER XIII

## CATHODE-RAY TUBES

**76. Fundamental Properties of Cathode-ray Tubes.**—The cathode-ray tubes used in communication work are always of the hot-cathode type and may be divided into high-vacuum and gas-filled tubes. A cross-section of a typical high-vacuum cathode-ray tube is shown in Fig. 159. The essential features are a heater cathode $C$ surrounded by a thimble-shaped control electrode $G$.

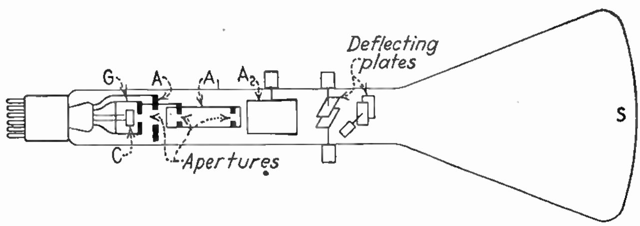



FIG. 159.—High-vacuum cathode-ray tube.

Next comes an accelerating electrode $A$ which is a disk quite close to the control electrode and is followed by two anode electrodes $A_1$ and $A_2$ operated at different potentials, the deflecting plates, and the fluorescent screen $S$. The entire tube is evacuated as thoroughly as possible following the usual vacuum-tube practice. The name cathode-ray tube comes from the fact that the tube depends for its operation upon a beam of electrons, or "cathode ray."

The assembly, consisting of the cathode, control electrode, accelerating electrode, and two anodes, is called an electron gun and is for the purpose of producing a high-velocity beam of moving electrons which when projected against the fluorescent screen will produce a small luminous spot.[1] A diagram showing the way in which the various parts of the electron gun are arranged and the way they are connected is shown in Fig. 160.

[1] For a discussion of the theory of the electron gun see I. G. Maloff and D. W. Epstein, Theory of Electron Gun, *Proc. I.R.E.*, vol. 22, p. 1386, December, 1934.

The control electrode is operated at a negative potential with respect to the cathode and serves the dual purpose of helping to concentrate the electrons into a beam, and of providing a means for controlling the number of electrons in the beam. The electrostatic field between the control electrode and the cathode is of such a character as to encourage the electrons to flow through the small aperture, with the number of electrons actually passing through the aperture limited by space-charge effects and controlled by the negative potential of the control electrode in relation to the positive potential of the accelerating electrode. The

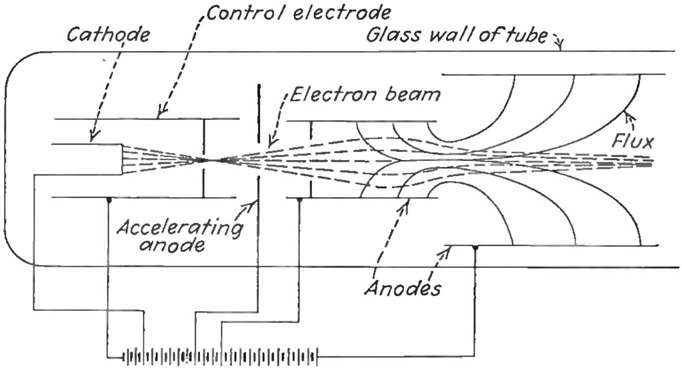



FIG. 160.—Circuit diagram of electron gun, together with electrostatic flux distribution that produces focusing, and the resulting electron beam.

strength of the electron beam is therefore controlled by the bias voltage of the control electrode in much the same way as the plate current of a vacuum tube is controlled by the bias on the grid.

The accelerating electrode is placed as close as possible to the control electrode and is operated at a moderate positive potential. It is provided with an aperture to intercept electrons which have widely divergent angles. The first anode accelerates the electrons further, and has one or more additional apertures to limit again the angle of divergence of the electron beam. In some cathode-ray tubes the accelerating anode electrode is eliminated and its function performed by the first anode. The second anode is operated at a positive potential much higher than the first anode, thereby providing further acceleration of the electrons, and is also so proportioned physically with respect to the first anode that the electrostatic fields produced between the two are such that the electrons in the cathode-ray beam focus into a very small spot at the fluorescent screen.

The electrostatic field that is between the two anodes has the general character shown in Fig. 160, and it can be shown that the result is equivalent to the effect which an ordinary thick lens has on light. This focusing effect is determined primarily by the relative potentials upon the two anodes and is not affected by the number of electrons within the beam, so that the intensity of the beam and hence of the spot at the fluorescent screen can be controlled by varying the control-electrode potential without otherwise affecting the behavior. A similar focusing action is also present in the region between the cathode and the control electrode. If no focusing means were provided, it would be necessary to make the apertures through which the electron beam must pass very small in order to keep the angle of divergence at a minimum, and this would mean relatively few electrons in the beam. Furthermore the electrons after passing the final anode would diverge as a result of their mutual repulsion, and the luminous spot would be large.

The second anode is sometimes formed by a silver coating on the inner walls of the tube extending from the fluorescent screen back to the first anode. This coating also eliminates the possibility of the glass walls accumulating electrostatic charges that would influence the cathode-ray beam.

*Gas-filled Cathode-ray Tubes.*—The physical structure of two common makes of gas-filled cathode-ray tubes is shown in Fig. 161, together with the circuits showing the way in which the electrodes are connected. The tube[1] of Fig. 161*a*, which is designed for medium anode voltage (1000 to 3000 volts), has a filament surrounded by a cylinder (often called a Wehnelt cylinder) which is operated at a fixed negative potential with respect to the cathode and is for the purpose of electrostatic focusing to reduce the size of the spot. The anode consists of a single plate having a small aperture which limits the angle of divergence within the beam. A final focus of the electrons within the beams is obtained by introducing traces of argon or other heavy inert gas within the tube.

[1] For further information on tubes of this type see Manfred von Ardenne, A Braun Tube for Direct Photographic Recording, *Exp. Wireless and Wireless Eng.*, vol. 7, p. 66, February, 1930.

The gas tube of Fig. 161*b* is designed for operation at anode voltages of the order of 250 volts.[1] The electron gun of this tube consists of a filament *F*, a diaphragm *D* operated at cathode potential and provided with a hole, followed by an anode consisting of a tube about 1 cm long and 0.1 cm in diameter, which insures that the emerging beam will be relatively concentrated. Final focusing to overcome the repulsions of the electrons within the beam during transit to the fluorescent screen is obtained by introducing a heavy inert gas at a pressure of 5 to 10 microns.

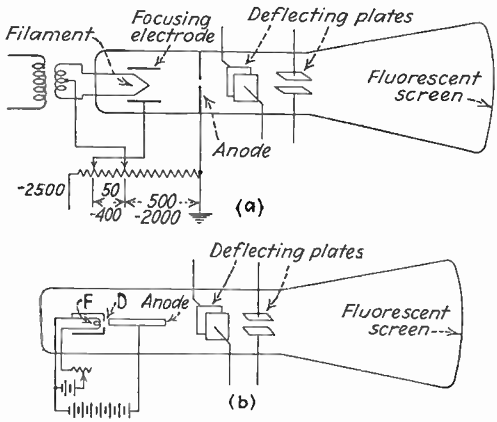



Fig. 161.—Medium- and low-voltage gas-filled type cathode-ray tubes.

The focusing action in a gas tube results from the fact that the electron beam ionizes the gas which it encounters in its path, giving rise to a positive space charge in this region because of the low mobility of the heavy positive ions. The result is the production of a radial electrostatic field about the cathode ray which draws the electrons together and thereby gives the possibility of obtaining a small spot on the fluorescent screen. The focusing action is controlled by the filament current since the higher the electron emission the greater will be the ionization and hence the stronger the radial field. All gas tubes must therefore be provided with a filament rheostat for focusing purposes, and they cannot have the intensity of the luminous spot controlled as can the high-vacuum tubes, because in attempting such control in a

[1] For further information on this tube see J. B. Johnson, The Cathode Ray Oscillograph, *Bell System Tech. Jour.*, vol. 11, p. 1, January, 1932.

gas tube one will disturb the focus and hence alter the spot diameter.

*The Luminous Screen.*—The light produced at the fluorescent screen depends upon the velocity and number of electrons in the cathode ray and the material of the screen. The most satisfactory material for visual observation is either natural or synthetic willemite, which has a luminous efficiency in lumens per watt approaching that of tungsten-filament lamps and produces a green light which has a wave length very near the point of maximum sensitivity of the eye. Furthermore willemite resists the burning that tends to occur as a result of the energy dissipated by the impinging electrons. The tendency of the light to persist after the beam has moved on depends on the screen material. With willemite, practically all visible luminescence ceases after about 0.06 sec., while with phosphorescent materials the persistence time may exceed 30 sec. The intensity of the spot is greater the higher the anode voltage. At voltages of the order of 200 to 500 volts the spot must be viewed in a subdued light and can be photographed only by allowing the spot to retrace its path over and over. In contrast to this, the spot produced at anode voltages from 2000 to 4000 volts is clearly visible in full daylight and a single rapidly moving trace is readily photographed.

The energy that is dissipated at the screen as a result of the impinging electrons depends upon the velocity and number of electrons, and is great enough in high-voltage tubes to burn the screen or even puncture the glass if a full intensity beam is allowed to be stationary for even a brief interval. It is hence necessary either to keep the spot in motion or to keep its intensity to a low value with the aid of a control electrode.

*Deflection of Beam.*—The beam that emerges from the electron gun may be deflected from its initial direction by the use of either electrostatic or magnetic fields. Tubes for general laboratory use are usually provided with two pairs of plates at right angles to each other, as shown in Figs. 159 and 161, which are for the purpose of deflecting the beam electrostatically. This deflection is given by the following formula:

$$D = \frac{1}{2}\frac{lLE}{dV} \tag{60}$$

where $V$ is the anode voltage, $E$ the voltage applied to the deflecting plates, and the remaining notation is as shown in Fig. 162.[1]

Examination of Eq. (60) shows that the deflection is a linear function of the deflecting voltage, and that for maximum deflection the plates should be long, close together, and as far as possible from the fluorescent screen, while the anode voltage should be low. The sensitivity to electrostatic deflection is hence limited by structural considerations, by the fact that the deflecting plates must not intercept the beam, and by the fact that with low anode voltages the intensity of the spot is small.

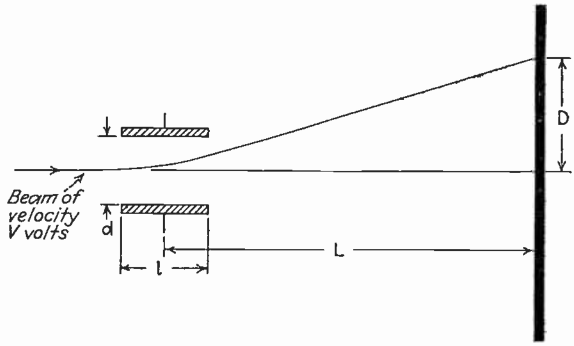



FIG. 162.—Deflection of electron beam by electrostatic fields.

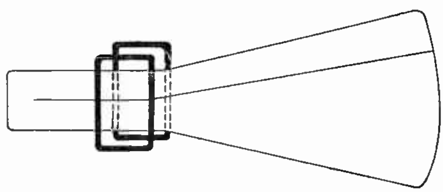
A magnetic field such that the lines of flux are at right angles to the direction in which the cathode-ray beam travels will cause a deflection in a direction at right angles to both the flux and the direction of the beam, as a result of the forces which are exerted upon a moving electron by a magnetic field. A suitable arrangement for producing magnetic deflection consists of two coils, as shown in Fig. 163*a*, placed on the screen side of the anode, just before the tube begins to flare out. The law of deflection with magnetic fields cannot be accurately given in a simple formula because of the varying strength of the magnetic field at

[1] This formula follows from the fact that the sidewise velocity which an electron acquires is proportional to the strength of the transverse or deflecting field, and the length of time which the electron is in this field. The deflection which results at the screen is proportional to the distance from the deflecting plates to the screen, and also inversely proportional to the velocity of the beam. The complete derivation is given by J. B. Johnson, The Cathode Ray Oscillograph, *Bell System Tech. Jour.*, vol. 11, p. 1, January, 1932.

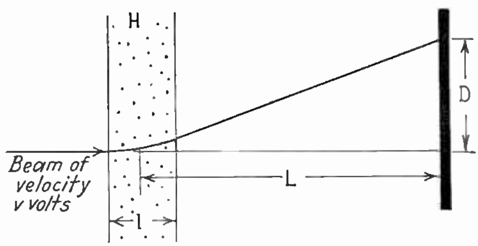
different positions along the beam path. If it could be assumed that the magnetic field was uniform over a limited distance $l$ and zero outside this region, then the deflection produced by a field strength $H$ would be given by the equation

$$D = \frac{0.3lLH}{V^{1/2}} \tag{61}$$

The notation is made clear by Fig. 163. It will be observed that the deflection is a linear function of the field strength,



(a) Arrangement of deflecting coils




(b) Idealized case of magnetic deflection

FIG. 163.—Arrangement of coils for magnetic deflection, and deflection of beam by magnetic flux.

and that maximum sensitivity is obtained when $l$ and $L$ are both large and the anode voltage is small.

Magnetic deflection provides a means whereby a current may deflect the cathode-ray beam, in contrast to electrostatic deflection in which the deflecting force is produced by a potential. Magnetic deflection is particularly useful where two deflecting forces must be applied at the same point, since then one may use either two pairs of deflecting coils at right angles to each other or one set of coils and one pair of electrostatic plates. From a practical point of view, magnetic deflection is limited to audio and the lower radio frequencies because of the difficulty of obtaining high-frequency fields of sufficient strength.

The frequency limit for electrostatic deflection is determined by the finite time required by the electrons to traverse the deflecting portion of the tube. Thus if the frequency of the voltage applied to the deflectors is so high that a complete cycle occurs while an electron is between the plates, the net result is zero deflection of the beam. The frequency limit in practical tubes is in the order of $10^8$ cycles, but varies somewhat with the size of the deflecting plates and the anode voltage. There is a similar limitation with respect to magnetic deflection, but this is not particularly important because of the difficulty of obtaining sufficient flux density at the higher radio frequencies.

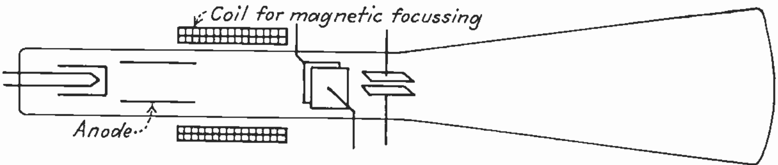



FIG. 164.—Cathode-ray tube with magnetic focusing coil and electrostatic deflection.

In gas-filled tubes the frequency limit is set by the fact that, if the beam travels too rapidly for ionization to build up to the proper value, the focusing action of the gas will be lost and the spot enlarged. This occurs at approximately 100,000 cycles. Another characteristic of gas-filled tubes using electrostatic deflection is a space charge between the deflecting plates that reduces the sensitivity to small deflecting voltages. This causes the calibration curve giving deflection as a function of deflecting potential to have a jog in it near the origin.

*Magnetic Focus.*—A magnetic field arranged so that flux lines are parallel to the direction in which it is desired that the electrons travel, as shown in Fig. 164, will serve to focus the electrons in the beam just as effectively as can be accomplished by electrostatic means or by gas. Such a field produces no force on electrons moving parallel with it; but if there is a velocity component at right angles to the field, the electron spirals along a helical path and ultimately returns to the beam. The point at which this return occurs is determined by the strength of the magnetic field in relation to the velocity of the beam in a direction parallel to the magnetic-flux lines, but is independent of the velocity which the electron has in a direction at right angles to the magnetic field. The result is therefore that all of the electrons having a radial component of velocity return to the beam at the same point, and by proper adjustment of the magnetic flux density this can be made the point at which the electrons impinge upon the fluorescent screen.[1] It is possible in this way to obtain a very sharp spot which is an exact image of the anode aperture through which the cathode-ray beam emerges. This method of focusing is particularly useful in conjunction with magnetic deflection, since then the focusing field can be made uniform through the path traversed by the cathode-ray beam. With electrostatic deflection it is necessary to localize the magnetic field as shown in Fig. 164 to prevent interaction between focusing and deflecting forces.

*Comparison of Gas-filled and High-vacuum Tubes.*—The gas-filled tubes are limited practically to low and medium voltages, while the high-vacuum type can be used at all anode voltages. It is found however that at low voltages the focus is best with gas-filled tubes, while at high anode potentials the vacuum type is superior. The gas-filled type has a relatively short life, often only 50 to 100 hr., whereas the high-vacuum type has the same life expectancy as an ordinary vacuum tube, *i.e.*, several thousand hours. Gas-filled tubes have the further disadvantage that they cannot be used at very high deflecting frequencies as mentioned above, and furthermore there is no simple means for modulating the beam intensity, since the focusing action is controlled by the number of electrons in the beam. The sensitivity to deflecting fields and the luminous intensity of the spot is a function only of the anode voltage, current in the beam, and focusing of the beam, and is not affected by the presence or absence of gas. The present trend is definitely toward high-voltage high-vacuum cathode-ray tubes because of their longer life and much higher luminous intensity as compared with medium- and low-voltage gas-filled tubes.

**77. Mounting and Installation of Cathode-ray Oscillograph Tubes.**—A cathode-ray oscillograph tube is handled in very much the same way as an ordinary vacuum tube except that special care must be taken to eliminate stray magnetic and electrostatic fields that would otherwise produce spurious

[1] For an analysis of this behavior see A. H. Brolly, Television by Electronic Methods, *Elec. Eng.*, vol. 53, p. 1153, August, 1934.

deflections. Proper attention to the orientation of near-by power transformers, filter reactors, etc., will eliminate spurious magnetic deflections from these sources. Electrostatic shielding is also desirable, but not always absolutely essential. Trouble from spurious deflecting forces is most pronounced in the case of low-voltage tubes because of their high sensitivity.[1]

The voltages required by the electrodes of a cathode-ray tube can be conveniently obtained from a very simple rectifier-filter system inasmuch as the total current drain of a cathode-ray tube seldom exceeds a few hundred microamperes. One can hence use either a half-wave rectifier tube shown in Fig. 165*a* or the voltage-doubling circuit of Fig. 165*b*, with a filter consisting of a single 2- to 4-$\mu$f condenser. A complete wiring diagram of a power-supply system suitable for use with high-vacuum cathode-ray tubes requiring 1000 and 2000 volts is given in Fig. 165. Transformer $T_1$ can be an ordinary replacement power transformer developing 350 to 375 r.m.s. volts on each side of the center tap. Because of the very low powers involved, all fixed resistors shown in the diagrams can be of the carbon composition type. The potentiometers, except $P_1$, can be ordinary untapered replacement volume controls. The output voltage developed is controlled by potentiometer $P_1$ which varies the voltage applied to the transformer primary. This potentiometer must have sufficient resistance and wattage capacity to enable it to be connected across the 110-volt line and handle the magnetizing current of the transformer primary in addition to its own bleeder current. The positive side of the circuit is grounded in order that the fluorescent-screen part of the tube may be at ground potential. This makes the filament of the cathode-ray tube, the control electrode, etc., at a high negative potential with respect to ground so that this part of the circuit should be inclosed, leaving only the fluorescent-screen end of the tube exposed. A separate transformer having adequate insulation between primary and secondary is used to supply filament power for rectifier and cathode-ray tubes. It is also possible to place additional windings on this filament transformer for

[1] In low-voltage tubes the earth's magnetic field is capable of producing appreciable deflection, as is also sheet iron that has become magnetized. Sheet iron used in the mounting must hence not be too close to the tube and in some cases may have to be demagnetized.

supplying a sweep circuit, and for biasing the spot to cut-off on the return sweep as discussed in Sec. 78. A separate transformer for this purpose is preferable, however. High-resistance leaks

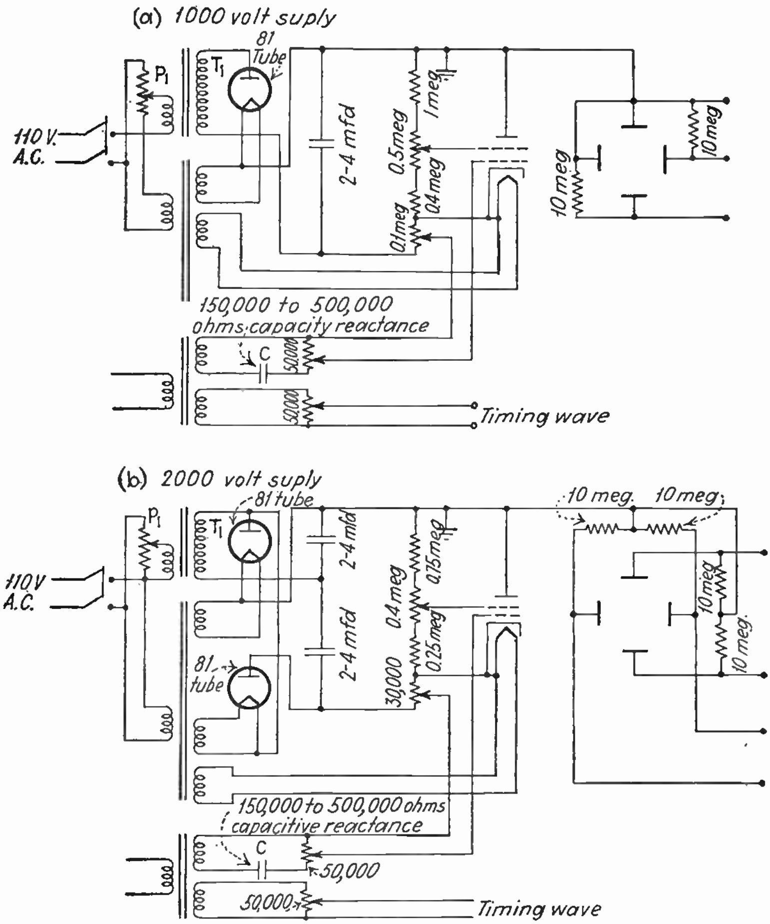



FIG. 165.—Power-supply systems for high-vacuum cathode-ray tubes.

are shown connected across each pair of plates, and the plates must be grounded either directly, or indirectly through a high resistance, in order to prevent high potentials from being built up as a result of stray electrons which the deflecting plates capture from the cathode-ray beam.

The arrangements shown in Fig. 165 can be simplified for many purposes by entirely omitting the transformer for supplying a timing wave and for turning off the spot during the return trace. The potentiometer $P_1$ can also be replaced by a switch that cuts in one or two steps of resistance in series with the transformer primary to reduce the anode voltage.

**78. Time Axis for Cathode-ray Tubes.**—In most uses to which a cathode-ray tube is put, the phenomenon to be observed is arranged to deflect the beam in one direction while a deflecting force that is a function of time is applied at right angles. This latter deflecting force is commonly referred to as the timing wave, or sweep wave, and may be either a sinusoidal alternating voltage

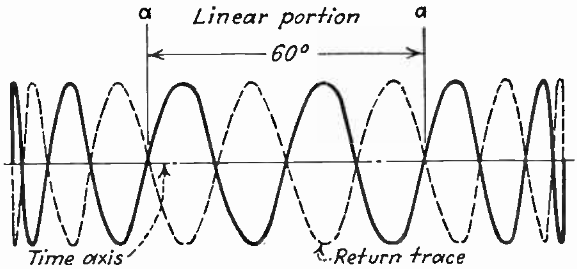



FIG. 166.—Typical pattern obtained with horizontal sinusoidal timing wave having one-twelfth the frequency of an alternating voltage applied to the vertical deflecting plates.

or a saw-toothed wave. By making the frequency of the timing wave a frequency that is $1/n$ of the frequency of the observed wave, when $n$ is an integer, a stationary pattern will result.

A sinusoidal timing wave gives a deflection along the timing axis that is proportional to sin $\omega t$, and so is a non-linear function of time as shown in Fig. 166. This disadvantage can be largely overcome, however, by using only the center half of the wave (the part between *aa* in Fig. 166), since this departs from linearity by only about $2\frac{1}{2}$ per cent. The amplitude of the timing wave can then be made so large that this useful part of the range covers the entire screen, while the non-linear portions strike the walls of the tube and are not seen. A sinusoidal timing wave also traces out a pattern on the return half cycle as shown dotted in Fig. 166. This return trace can be prevented from appearing on the screen in high-vacuum tubes by applying to the control electrode a voltage having the same frequency as the timing voltage, but approximately 90° out of phase with it. In this way the return trace can be diminished in intensity, or even made to disappear entirely.

A sinusoidal timing voltage, and also the potential necessary to suppress the spot during the return half cycle, can be conveniently applied to the cathode-ray tube through a well-insulated transformer having two secondaries as shown in

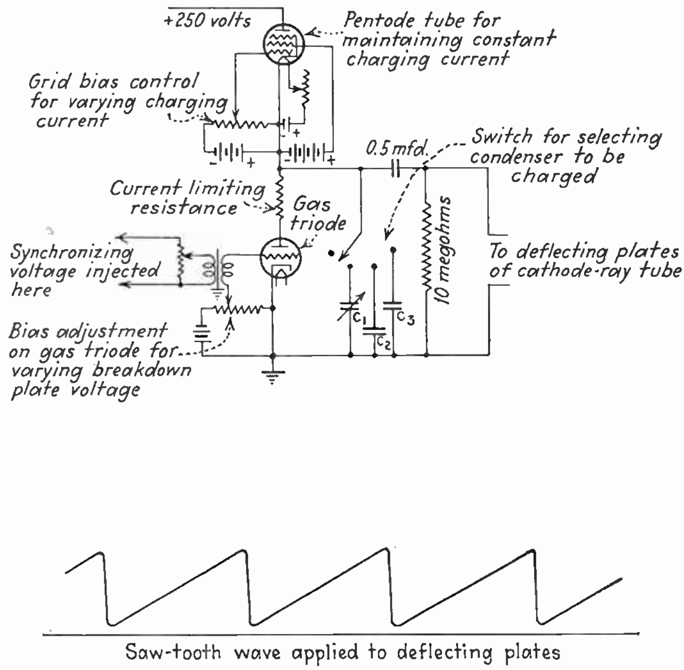



FIG. 167.—Typical saw-tooth oscillator, with saw-tooth wave obtained at the output terminals.

Fig. 165. The primary of this transformer can be connected to the 110-volt 60-cycle power source when a 60-cycle sweep is satisfactory, or can be switched to a laboratory oscillator if a different sweep frequency is desired.

The most desirable timing wave has a saw-tooth shape as shown in Fig. 167. Such a wave causes the spot to travel across the screen in one direction at a constant rate, giving a linear time scale, and jump back almost instantly across the screen to retrace its path. Saw-tooth waves are obtained by charging a condenser at a constant rate and then periodically discharging through a grid-controlled gas triode.[1] A typical saw-tooth oscillator circuit is shown in Fig. 167. The condenser charging current is made constant by taking advantage of the fact that the plate current of a radio-frequency pentode tube is substantially constant provided the plate voltage is at least 15 per cent of the screen potential. The magnitude of the charging current can be controlled by varying the grid bias of the pentode tube. As the condenser receives charge through the tube, its potential builds up as a linear function of time until sufficient to start current flow in the gas tube, at which instant the condenser discharges almost instantly through the gas tube. It might be thought that once the gas tube becomes conducting it would continue in this condition, but this is not the case because the inductance of the connecting wire causes the discharge to overshoot slightly, reducing the voltage across the gas tube below the point at which ionization can be maintained, and thereby giving the deionization process a brief instant to take hold. The result is that the tube becomes non-conducting immediately after the discharge, and the cycle repeats.

The amplitude of oscillations is controlled by the breakdown potential of the gas tube, which is determined by the grid bias of the tube. The maximum amplitude obtainable is fixed at low frequencies by the peak plate-voltage rating of the gas tube, and at higher frequencies by the fact that time must be allowed for the deionization process to be able to get started. This can be explained with reference to Fig. 168*a*, where it is seen that the larger the amplitude of oscillations and the higher the frequency, the faster the voltage will rise after the initial discharge. The deionization process requires a brief interval marked $T_1$ to become complete, and if the voltage rises to too high a value in this period the gas tube will reionize, discharging the condenser again. The result is then as shown at Fig. 168*c* when the voltage rises

[1] Such tubes are sold under various names such as thyratron, grid-glow, trion, gas-triode. The outstanding characteristic of these tubes is that, if the grid potential is less than cut-off even momentarily, positive ions are produced in the tube which neutralize the negative space charge of the electrons and also the controlling action of the grid, with the result that the voltage drop across the tube drops to somewhere between 10 and 20 volts.

much too rapidly, and as at Fig. 168*b* when it rises only slightly faster than permissible. At high frequencies one must hence limit the amplitude of oscillation by increasing amounts as the frequency of the saw-toothed oscillator is increased, and amplification may become necessary to give sufficient deflection.

It is desirable that the saw-toothed oscillator be able to maintain its straight-line character even during the deionization

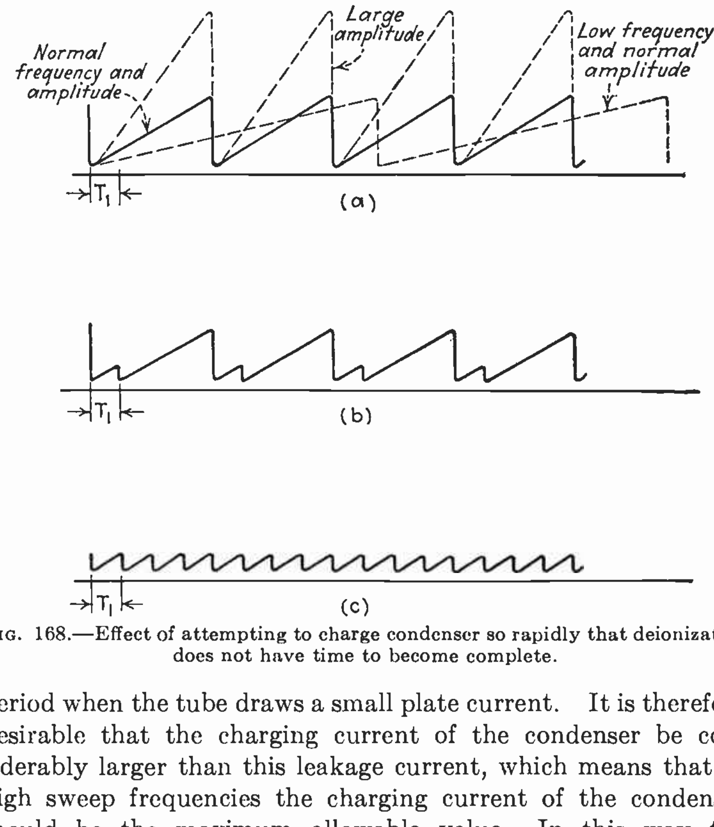



FIG. 168.—Effect of attempting to charge condenser so rapidly that deionization does not have time to become complete.

period when the tube draws a small plate current. It is therefore desirable that the charging current of the condenser be considerably larger than this leakage current, which means that at high sweep frequencies the charging current of the condenser should be the maximum allowable value. In this way the straight-line part of the cycle is lengthened as shown in Fig. 169, and the useful amplitude is increased. At low frequencies the deionization time is negligible compared with the duration of a cycle and so does not need to be considered. The smallest permissible value of charging current is then determined by the fact that the charging current must be at least ten times the current drawn by the circuit to which the saw-tooth wave is applied. This is because the current drawn by this circuit is proportional to voltage instead of being constant, and thereby tends to distort the saw-tooth wave. This places the minimum allowable charging current at a rather high value, as seen from the following example: Assume that the total voltage variation across the plate of the gas triode is 200 volts and that the deflecting plates to which the saw-tooth wave is applied are shunted by 10 megohms.

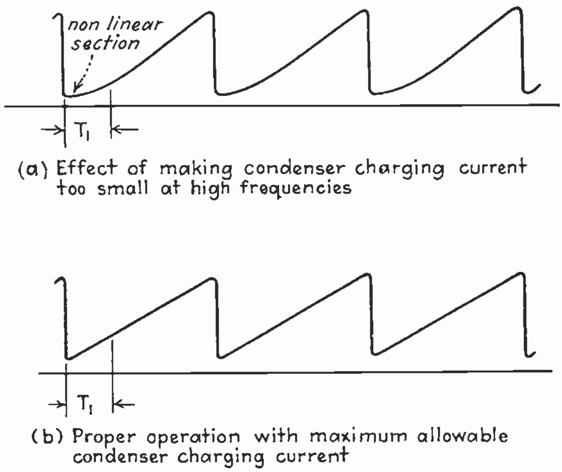



FIG. 169.—Effect of different condenser charging currents when generating high-frequency oscillations. A small charging current makes the wave curved during the deionization period $T_1$, thereby reducing the useful (*i.e.*, linear) part of the oscillation.

Then the maximum current flowing into the 10 megohm is $100/10^7 = 10$ $\mu$a so that the condenser charging current must be not less than 0.1 ma if it is to be ten times as great. It is thus apparent that the circuit to which the saw-tooth voltage is applied should have a very high resistance.

The frequency of the saw-toothed oscillations is determined by the time required for the charging current $I$ flowing into the capacity $C$ to increase the potential across this condenser from about 15 volts to the plate potential $E_p$ at which the gas tube will break down. Consequently

$$\text{Frequency in cycles per second} = \frac{I}{C(E_p - 15)} \tag{62}$$

where $C$ is in farads, $I$ in amperes, and $E_p$ in volts. When high frequencies are generated, the operating procedure is to set the charging current to the maximum allowable value, and then to control the frequency by varying the capacity $C$ while using the highest grid bias on the gas tube (which means the highest breakdown plate voltage) that will allow deionization to take place properly. In generating low frequencies the grid bias of the gas tube is set to give the breakdown plate voltage corresponding to the amplitude desired, the capacity is set at an appropriate large value, and the final adjustment of frequency obtained by varying the charging current of the condenser, keeping in mind that the charging current should never be allowed to fall below some limiting minimum value. The problem of making the sweep circuit operate properly at any desired frequency can be simplified by preparing a curve giving the proper grid-bias setting of the gas tube as a function of frequency, and indicating when the frequency is controlled by keeping the charging current at a maximum and varying the capacity, and, when it is controlled by setting the capacity at a maximum and varying the charging current.

The gas tube introduces certain limitations other than those involving the deionization time. Any particular tube has a maximum peak plate-voltage rating which cannot be exceeded without danger of flashback in the tube, and also each tube has a peak-current rating which cannot be exceeded without damaging the cathode. In order that this peak current will not be exceeded during the discharge period, it is necessary to place in series with the plate of the gas tube a resistance equal to the peak-voltage rating divided by the allowable peak current. The performance at higher frequencies depends largely upon the molecular weight of the gas within the tube, since heavy ions require more time for deionization than do the more mobile, lighter ions. As a result, mercury-vapor gas tubes are not satisfactory above a few thousand cycles, whereas tubes filled with gases such as argon will operate at considerably higher frequencies.

Complete circuit diagrams for the production of a saw-tooth wave for cathode-ray oscillograph purposes are shown in Figs. 167 and 170. The principal problem involved in these circuits arises from the fact that the cathodes of the pentode and gas tubes cannot both be at ground potential. One solution is to use a dry-cell type of pentode tube and supply it with separate filament, screen, and bias batteries as shown in Fig. 167. An alternative arrangement is shown in Fig. 170, where an alternating-current power-supply arrangement is used to furnish all voltages except the grid bias for the gas tube. With an arrangement of this type, considerable care must be taken to prevent spurious alternating voltages being developed across the output of the oscillator. In particular, the power transformer must have a grounded electrostatic shield between

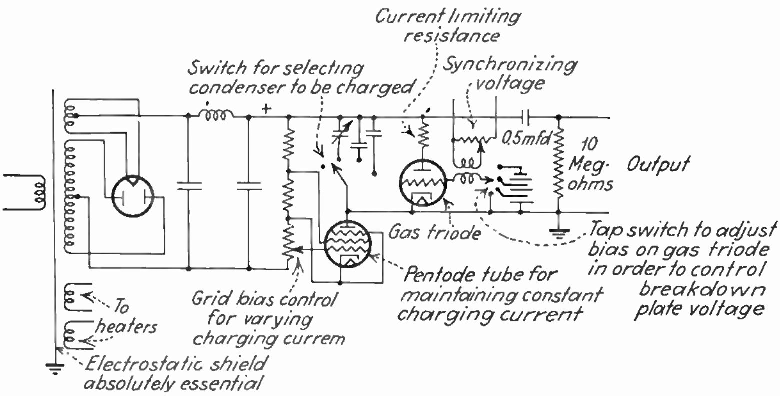



FIG. 170.—Saw-tooth oscillator circuit operating from 60-cycle power.

primary and secondary. The pentode and gas tubes should also be of the heater type if possible, as otherwise separate filament transformers having electrostatic shields between primary and secondary are required.

The frequency of oscillation of a saw-tooth oscillator can be readily controlled by injecting a small voltage into the grid circuit of the gas tube, in very much the same way that the frequency of a multivibrator can be controlled. This action results from the fact that when the grid bias of the tube contains an alternating component, the plate potential at which breakdown occurs varies as shown in Fig. 171, so that breakdown tends to occur at regular intervals controlled by the injected oscillation. It is possible in this way to make the sweep circuit have a frequency $1/n$ of the injected frequency, where $n$ is an integer. The stability of the control is greatest when the uncontrolled frequency of oscillation is very slightly lower than the desired frequency, and it is possible with proper adjustments to obtain stable operation when the ratio of injected to saw-tooth frequency is as great as 10 to 1. The proper procedure for synchronizing the sweep circuit with an injected voltage is to vary the uncontrolled frequency until it is very nearly correct, and then to inject the smallest possible synchronizing voltage that will give stable control. If the synchronizing voltage is increased in amplitude beyond this point, the frequency will increase in discontinuous jumps, and the amplitude of oscillations will be likewise decreased, as is apparent from a study of Fig. 171.

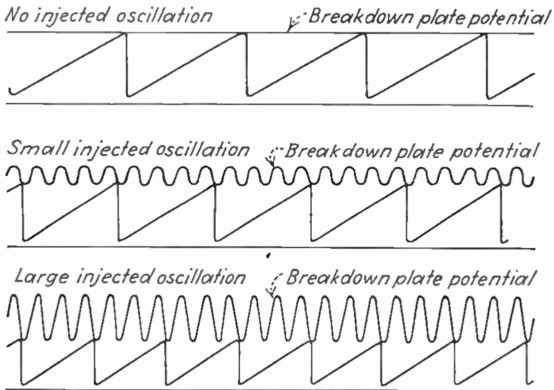



FIG. 171.—Oscillograms showing mechanism by which saw-tooth frequency can be controlled by an injected frequency.

When the voltage used to deflect the spot vertically and the voltage injected into the saw-tooth oscillator are derived from a common source, there is danger of distortion being produced in the observed pattern as a result of the reaction exerted by the sweep circuit, partly as a result of the grid current in the gas triode, and partly as a result of capacitive couplings. Trouble of this sort can be avoided by introducing the synchronizing voltage into the grid circuit of the gas triode through a shielded transformer with suitable turn ratio, and by proper arrangement of parts.

**79. Cathode-ray Tube Applications.**—The following paragraphs describe some of the more common uses of cathode-ray tubes and give an idea as to the possibilities of such tubes.[1]

[1] A more complete discussion of the subject is given by R. A. Watson-Watt, "The Cathode-ray Oscillograph in Radio Research," His Majesty's Stationery Office, London.

In many of the cathode-ray applications considered below it is necessary to measure deflections. This can be done with high accuracy by pasting a piece of transparent cross-section paper over the fluorescent screen. Under some circumstances it may even be desirable to place calibration lines directly on the glass.

*Measurement of Voltage and Current.*—A cathode-ray tube can be used as a voltmeter by applying the voltage to be measured to a pair of electrostatic deflecting plates and measuring the resulting deflection. The tube acts as an electrostatic voltmeter in that it consumes practically no energy, and it measures the crest amplitude of an alternating wave at all radio frequencies. Calibration can be obtained by the use of known direct-current voltages, and the sensitivity obtainable with ordinary tubes is in the range 1.0 to 0.2 mm deflection per volt.

A cathode-ray tube can be used as an ammeter by passing the current to be determined through magnetic deflecting coils and measuring the resulting displacement of the cathode-ray beam. This gives the crest value of the alternating current at all radio frequencies. A calibration can be made with direct current, and the sensitivity when the deflecting coils are close to the neck of the tube is of the order of 1.0 to 0.1 mm deflection per ampere turn. At very high frequencies it may be necessary to take into account the fact that the distributed capacity of the coil gives rise to a circulating current that makes the actual current flowing in the inductance slightly larger than the current being measured.

*The Cathode-ray Tube as an Oscillograph.*—A very important use of the cathode-ray tube is as an oscillograph to observe wave shapes. Oscillograph operation is obtained by making the horizontal deflection a function of time while using the phenomenon being observed to supply the vertical deflection. Where possible, it is desirable that the deflection along the horizontal be made proportional to time either by the use of a saw-tooth oscillator or by using selected portions of a sinusoidal wave as discussed in Sec. 78. When this is not feasible, a sinusoidal timing wave such as illustrated in Fig. 166 is used.

When periodic phenomena are observed, it is desirable where possible to make the timing wave have a frequency that is exactly $1/n$ of the frequency of the periodic wave, where $n$ is an integer preferably not 1. In this way a stationary pattern will be obtained upon the fluorescent screen; and if $n$ is larger than 1, several cycles of the wave being investigated will be obtained.

The case where the timing wave is sinusoidal, and of the same frequency as the wave under investigation, is of sufficient importance to warrant individual consideration. When both

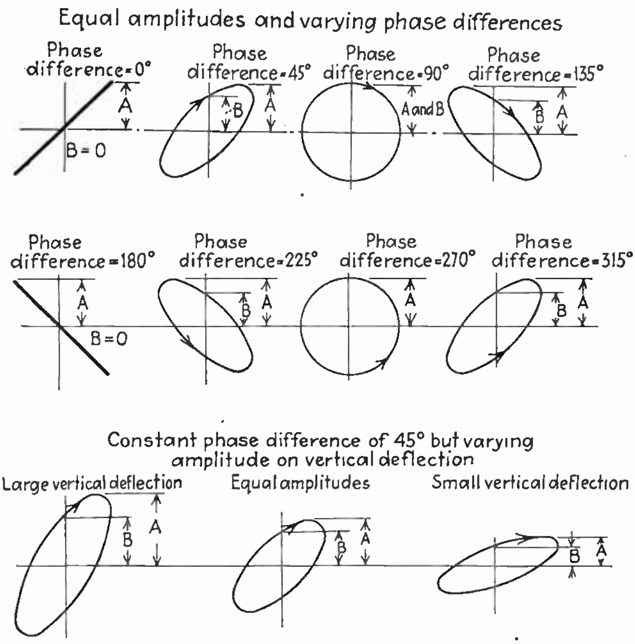



FIG. 172.—Typical patterns produced by cathode-ray tube when sinusoidal forces of the same frequency but differing phase and amplitudes are used to deflect the spot.

horizontal deflection and vertical deflections are the result of sinusoidal forces of the same frequency, the resulting pattern is an ellipse, the character of which depends upon the relative amplitudes and phases of the two deflecting forces. Typical illustrations are shown in Fig. 172 where it will be seen that when the two voltages have the same phase the ellipse degenerates into a sloping line, while when the phase difference is exactly 90° and the two amplitudes are equal the ellipse becomes a circle.

The phase difference $\alpha$ between the two deflecting forces can be readily determined from the relation

$$\sin \alpha = \pm \frac{B}{A} \tag{63}$$

The notation is as in Fig. 172, and the quadrant must be worked out from the orientation of the major axis of the ellipse and the direction in which the spot travels. Uncertainty as to the direction in which the spot travels can always be eliminated by shifting the phase of one of the deflecting voltages in a known direction and noting the effect on the pattern. Thus if in Fig. 172 the pattern is such that the phase difference is either 45 or 315°, adding 45° to the shift will give a circle for 45° shift, and a

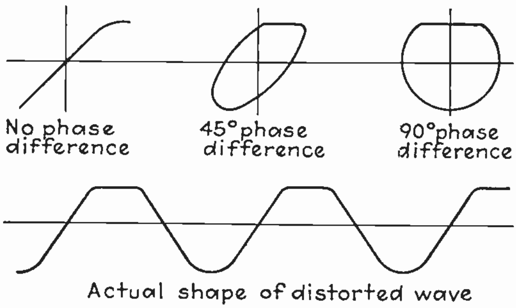



FIG. 173.—Typical patterns produced by cathode-ray tube when a sinusoidal deflecting force acts along the horizontal axis and a distorted sinusoidal force of the same frequency acts along the vertical axis, together with oscillograms of distorted wave plotted on a linear time axis.

straight line for 315° shift.[1] The cathode-ray tube furnishes the simplest means available for determining phase differences at radio frequencies.

When one wave is sinusoidal and the other is a distorted wave of the same frequency, the resulting pattern is a distorted ellipse as shown in Fig. 173. The actual wave shape can be readily derived by replotting the wave on a linear time axis as shown.

The cathode-ray oscillograph possesses none of the frequency limitations of the ordinary oscillograph. The pattern produced on the fluorescent screen by periodic phenomena can be readily photographed by using a suitable time exposure. With high-voltage cathode-ray tubes the intensity of the spot is sufficiently great to permit photographing a single trace of the fluorescent

[1] Another method for determining the direction of travel of the spot is described by E. R. Mann, A Device for Showing the Direction of Motion of the Oscillograph Spot, *Rev. Sci. Inst.*, vol. 5, p. 214, July, 1934.

spot, making it possible to photograph transients through the audio-frequency range.[1]

*Measurement of Power.*—The cathode-ray tube can be used to measure power by means of the circuit shown in Fig. 174. The deflection along one axis is made proportional to the instantaneous voltage, while along the other axis it is made proportional to the instantaneous integral of the current, *i.e.*, to the charge

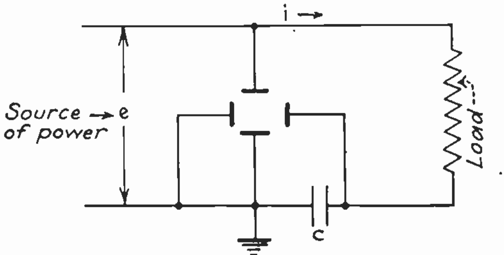



FIG. 174.—Cathode-ray tube connected to measure power dissipated in a load. The area of the pattern traced by the luminous spot is proportional to power, provided the losses in the condenser $C$ are negligible.

upon the capacity $C$. The resulting diagram traced by the fluorescent spot is a closed-area figure in which the area is proportional to the power loss per cycle and hence to the average power.[2] Calibration can be made by introducing known losses or by calculating the relation between area and power from the measured sensitivity of the instrument. The cathode-ray tube can be used in this way to measure accurately the power dissipated in a circuit at radio frequency. The only error that can occur is from losses in the condenser $C$. The chief disadvantage of the method is that it is necessary to measure the area of what may be an irregular figure by the inconvenient

[1] An ingenious method of supplying a time axis for certain types of transients is described by E. V. Sundt and G. H. Fett, A Timing Method for Cathode-ray Oscillographs, *Rev. Sci. Inst.*, vol. 5, p. 402, November, 1934.

The transient is applied to one pair of deflecting plates, while the spot is interrupted by applying an audio frequency such as 5000 cycles to the control electrode of the cathode-ray tube. The pattern then consists of a line of spots spaced 0.0002 sec. apart for the 5000-cycle case, and it is a simple matter to replot on an amplitude-time coordinate system.

[2] This method was devised by Harris J. Ryan. See his paper, A Power Diagram Indicator for High Tension Circuits, *Trans. A.I.E.E.*, vol. 30, pt. II, p. 1089, 1911, for derivation of the equations involved.

process of placing transparent cross-section paper over the end of the tube and counting squares, or by use of a planimeter.

*Comparison of Frequencies.*[1]—An important application of the cathode-ray tube is in comparing frequencies which are to be adjusted in harmonic relationship to each other. When a sine wave of one frequency deflects the cathode-ray spot along one axis and a sine wave of a second frequency produces a deflection at right angles, then when the ratio of the two frequencies can

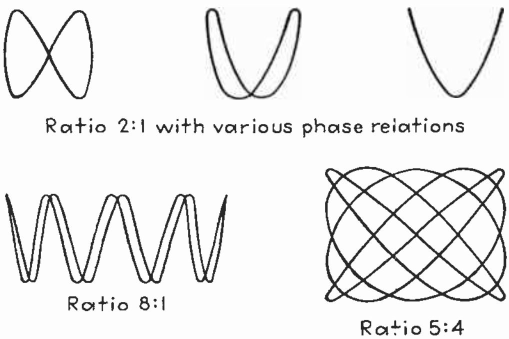


FIG. 175.—Typical Lissajous figures.

be expressed by an integer or a ratio of simple integers the result is a simple pattern called a Lissajous figure. The exact configuration of the pattern depends upon the frequency ratio and upon the relative phase of the two waves and is shown in Fig. 175 for typical cases. It is to be noted that the elliptical traces shown in Fig. 172 are Lissajous figures for the special case in which the frequency ratio is unity.

When the ratio of frequencies is an exact integer, the pattern is stationary and the frequency ratio is the number of times the side of the figure is tangent to a horizontal line divided by the number of times its end is tangent to a vertical line.[2] If the ratio is nearly but not exactly an integer, then the pattern weaves about as though the relative phase of the two deflecting waves was continuously changing, while if the ratio of frequencies

[1] For more information on this subject see F. J. Rasmussen, Frequency Measurements with the Cathode-ray Oscillograph, *Trans. A.I.E.E.*, vol. 45, p. 1256, 1926.

[2] This assumes that the forward and return trace do not coincide. If they do coincide, the rule does not necessarily hold.

differs very much from a simple ratio of integers the pattern is merely a luminous area.

When the frequency ratio is large, the pattern becomes so complicated that it is difficult to determine what the exact ratio

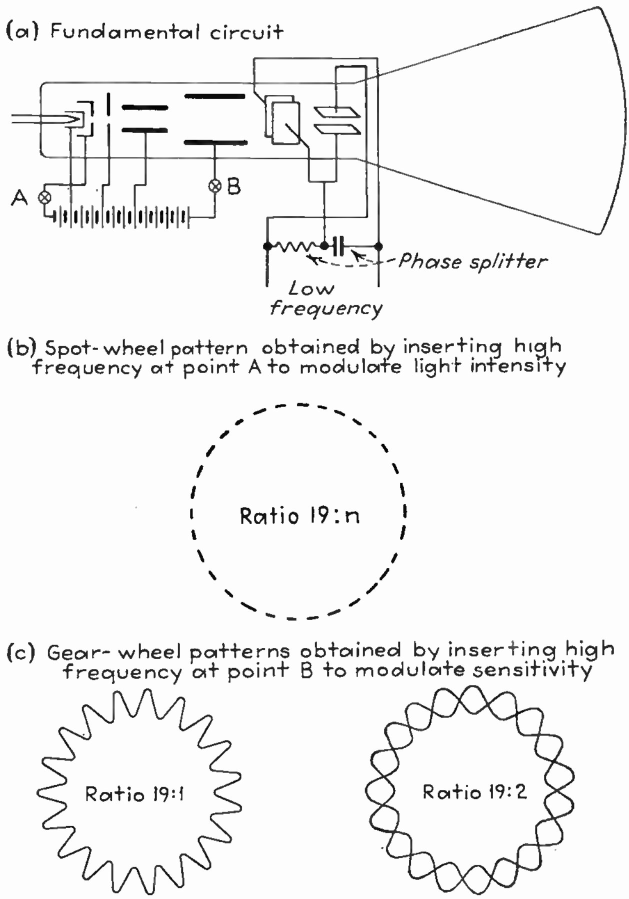



Fig. 176.—Wheel patterns used to compare frequencies.

is by inspection. In such cases arrangements such as shown in Fig. 176 can be used. Here the arrangements are such that the low frequency is caused to produce an elliptical or circular path by use of a resistance-capacity phase splitter. In Fig. 176*b* the beam intensity is modulated by the high frequency, giving a pattern as shown, with the ratio of high to low frequency equal to the number of spots divided by some integer. At Fig. 176*c* the high frequency is inserted in series with the anode. This varies the deflection produced by the low-frequency potentials at the deflecting plates because the higher the anode voltage the less the sensitivity of the tube. The result is a gear-wheel pattern with the ratio of the high to low frequency determined by the number of teeth. In Fig. 176 the wheel or the spots, as the case may be, appear to rotate if the frequency ratio is almost but not exactly expressible as a ratio of simple integers.

*Amplitude Modulation.*—The cathode-ray tube can be used in a number of ways to determine the degree and linearity of

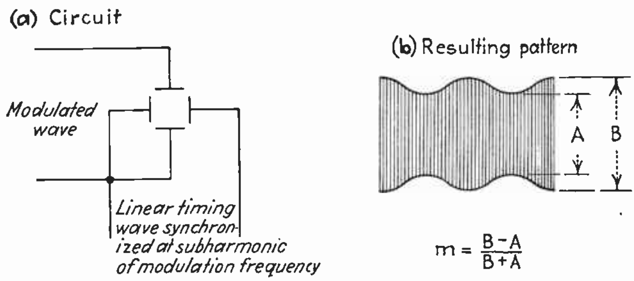



FIG. 177.—Modulated wave envelope obtained by use of linear sweep circuit synchronized at half the modulating frequency.

modulation. An obvious arrangement is to use a linear time axis synchronized at a subharmonic of the modulation frequency, and to apply the modulated wave directly to the other axis. The result is as shown in Fig. 177 and outlines the envelope of the modulated wave so that the degree of modulation and the presence of distortion can be easily determined. In a pattern such as shown in Fig. 177 the carrier frequency is usually so high compared with the sweep frequency that the vertical traces of successive cycles overlap to give a solid pattern.

Another common method of determining modulation is to apply the modulating wave to one pair of deflecting plates and the modulated voltage to the other pair. The resulting pattern is of the type shown in Fig. 178, and the degree of modulation can be determined from the formula given in the figure. The top and bottom parts of the pattern are straight lines when the modulation is linear, *provided* there is no phase shift between the modulating voltage applied to the cathode-ray tube and the modulation envelope. If the top and bottom sides of the pattern are curved instead of being straight lines, distortion is present, or there is a phase shift between the modulation envelope and the modulating voltage. The latter effect, if present, can be eliminated by shifting the phase of the modulating voltage

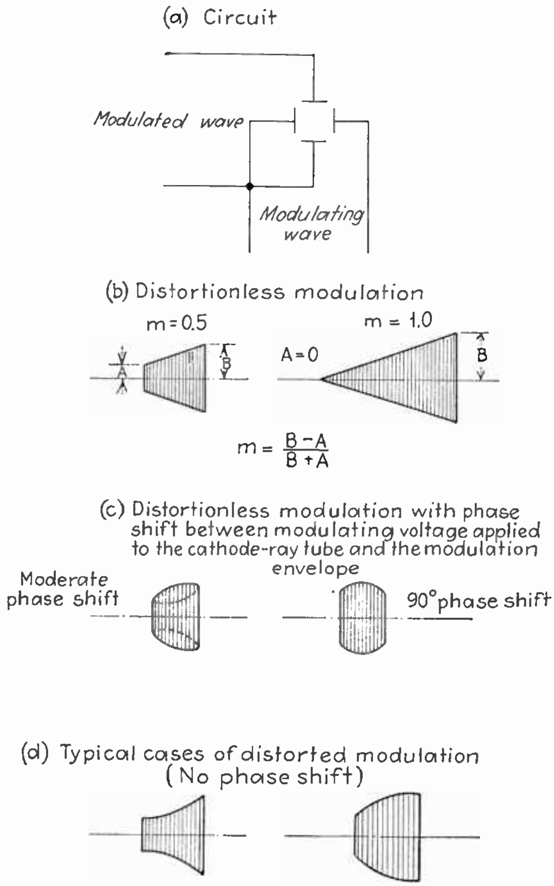



FIG. 178.—Circuit arrangement for observing modulation, together with typical resulting patterns.

applied to the cathode-ray tube. When no phase shift is present, this method of observation can be used when the modulation is non-sinusoidal or even non-periodic.

Another possibility is to rectify the modulated wave and apply the result to one pair of deflecting plates while using the audio-frequency modulating voltage to supply a timing axis. The result will be either a straight line or an ellipse if no distortion is present, but will be a curved line or a distorted ellipse if the modulation is not linear.

A continuous observation of the degree of modulation for monitoring purposes can be obtained by applying the modulated wave to one pair of deflecting plates and short-circuiting the second pair of plates. The resulting trace will be a straight line having an amplitude equal to the crest amplitude of the wave at the peak of the modulation cycle. Hence with 100 per cent modulation the line will be twice as long as it is when there is no modulation, and any excess in length over that corresponding to 100 per cent modulation is a sign of over modulation.

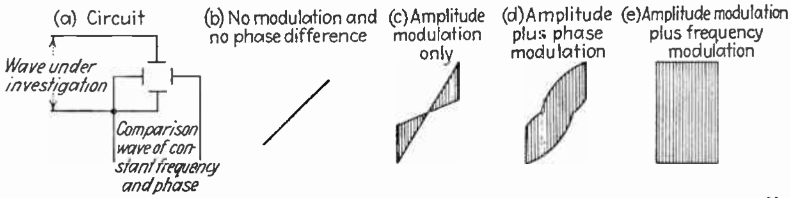



FIG. 179.—Detection of phase and frequency modulation accompanying amplitude modulation.

*Phase and Frequency Modulation.*—The cathode-ray tube supplies almost the only simple means of detecting the presence of phase or frequency modulation in amplitude-modulated waves. The procedure is to apply to one pair of deflecting plates a wave of the carrier frequency free of both phase and frequency modulation, while applying to the second pair of deflecting plates the wave being investigated. When no modulation of any sort is present, the resulting pattern is either a straight line or an ellipse, and if an ellipse it can be made a straight line by suitably shifting the phase of the voltage applied to one pair of plates. Assuming that matters have been arranged so that a straight-line pattern is obtained, then amplitude modulation free of phase and frequency modulation will produce a luminous area as shown at Fig. 179*c*. If, however, the amplitude modulation is accompanied by phase modulation, the pattern is as shown at Fig. 179*d*, being bounded by curved lines. If the amplitude modulation is accompanied by frequency modulation, the pattern then changes as shown in Fig. 179*e*.

The principal difficulty encountered in detecting frequency and phase modulation is in obtaining a suitable timing axis, since this must have exactly the same frequency as the carrier even to the extent of maintaining constant phase position. In the case of modulated amplifiers separated from the oscillator by adequate buffer stages, the timing wave can be obtained from the oscillator, but in the case of modulated oscillators it is necessary to employ a crystal oscillator for the timing wave. By careful adjustment it is possible in this way to maintain substantial constant phase and frequency over time intervals of a few seconds' duration.

The procedure outlined above determines only the presence of frequency and phase modulation. No simple means has yet

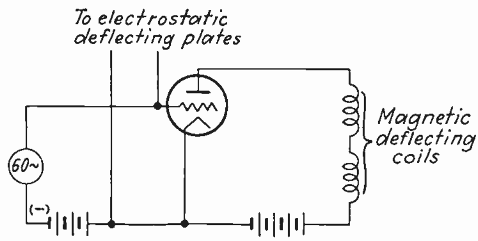



Fig. 180.—Circuit arrangement for obtaining $E_g - I_p$ curve of triode with a cathode-ray tube.

been devised for measuring quantitatively the extent of phase and frequency modulation accompanying amplitude modulation other than by replacing the modulating voltage by an adjustable direct-current potential in place of the usual modulating voltage, and then observing point by point the variation in phase and frequency.

*Tube Characteristics.*—Characteristic curves of tubes, such as the relation between plate current and grid voltage for constant plate potential, can be traced out by a cathode-ray tube. Typical circuit arrangements are shown in Fig. 180, where the vertical deflection is obtained by magnetic deflecting coils through which the plate current passes, while the horizontal deflection is the potential applied to the grid and consists of a negative bias voltage plus a 60-cycle component. Curves showing plate current as a function of plate voltage for constant grid potential can be obtained in similar manner. The cathode-ray tube is particularly useful in determining tube characteristics in the positive grid region where curves can not be taken point by point in the usual manner because of the excessive heating.

*Amplifier Characteristics.*—Distortion and phase shift in amplifiers can be determined by applying to one pair of deflecting plates a potential having the same phase as the amplifier input, but much greater amplitude, while using the amplifier output to give deflection along the other axis. If there is no distortion and no phase shift, the resulting pattern will be a straight line. Phase shift between input and output voltages produces an elliptical pattern from which the amount of phase shift can be determined as described above. Distortion causes either a bent straight line or a distorted ellipse.[1]

The dynamic characteristic of a vacuum-tube amplifier, *i.e.*, the relationship between instantaneous plate current and instantaneous plate voltage, can be determined by using a technique similar to that described above for obtaining tube characteristics. The instantaneous plate voltage operates one set of deflecting plates while deflection along the other axis is obtained either by the use of magnetic deflecting coils or by utilizing the voltage drop of a resistance in series with the plate.

*Visuals.*—Visuals are devices for giving a visual picture of a resonance curve of an amplifier, radio receiver, etc. The most satisfactory devices of this sort make use of cathode-ray tubes, and a typical arrangement is shown schematically in Fig. 181.[2] The tuned circuit or the receiver under investigation is excited by an oscillator which has its frequency varied continuously over the appropriate range by the use of a motor-driven rotating condenser. The output of the circuit or receiver under test is applied to one pair of deflecting plates of a cathode-ray tube, and amplification is used if this is required to give sufficient deflection. The timing wave applied to the other axis consists of a saw-tooth voltage controlled by a contactor on the motor shaft. This contactor is so arranged that it allows the sweep voltage to increase linearly with time during the period the condenser capacity varies from maximum to minimum, and then keeps the gas triode in a conducting condition during the half revolution in which the condenser returns from minimum to

[1] See F. C. Willis and L. E. Melhuish, Load Carrying Capacity of Amplifiers, *Bell System Tech. Jour.*, vol. 5, p. 573, October, 1926.

[2] For constructional information on this equipment see R.C.A. Application Note, No. 43, "Cathode-ray Curve-tracing Apparatus for Aligning Tuned Circuits," obtainable from the R.C.A. Radiotron Co., Harrison, N. J.

maximum capacity. The resulting trace is as shown in Fig. 181*b* and gives response as a function of frequency. A modification involves rectifying the radio-frequency voltage before applying it to the deflecting plates, and gives the result shown in Fig. 181*c*.

Visual test devices are used by some manufacturers to align radio receivers. These devices are particularly useful in high-

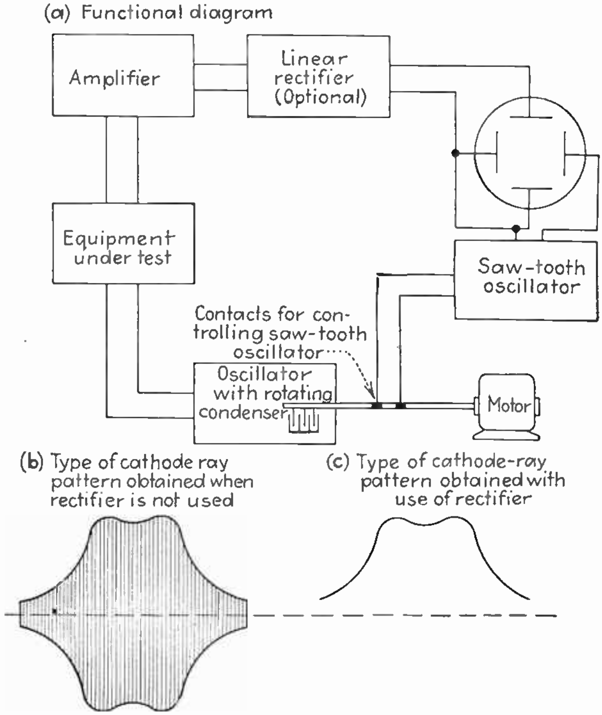



FIG. 181.—Functional diagram of visual test device, together with types of patterns that can be produced.

quality receivers employing band-pass filters since the visual indication greatly facilitates the obtaining of square-topped resonance curves.

*Cathode-ray Tubes in Investigation of Wave Propagation.*—The cathode-ray tube furnishes several excellent means of observing pulse signals in Kennelly-Heaviside-layer experiments. One arrangement suitable for this purpose is illustrated in Fig. 182.[1]

[1] See J. P. Schafer and W. M. Goodall, Kennelly-Heaviside Layer Studies Employing a Rapid Method of Virtual-height Determination, *Proc. I.R.E.*, vol. 20, p. 1131, July, 1932.

The received pulses are applied to one pair of deflectors while a linear timing wave is applied to the other. The pulses are synchronized with the 60-cycle power frequency so that one pulse per cycle is transmitted. The sweep circuit is also synchronized with the same power system so that successive pulses of the same retardation will superimpose and give a stationary pattern as in Fig. 182. The time delay is proportional to the corresponding distances measured along the linear time axis from the ground-wave pulse.

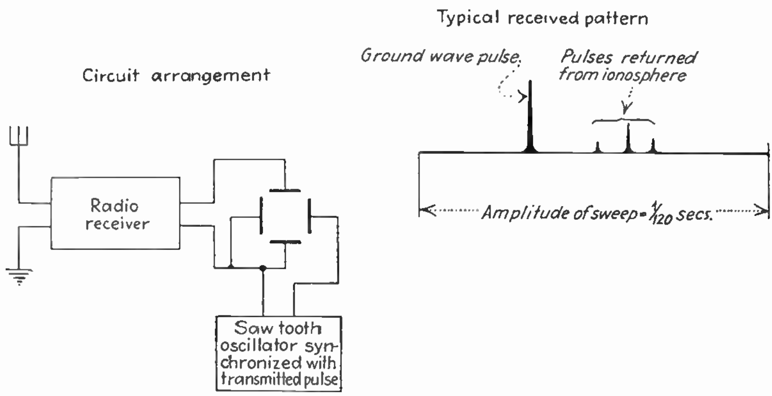
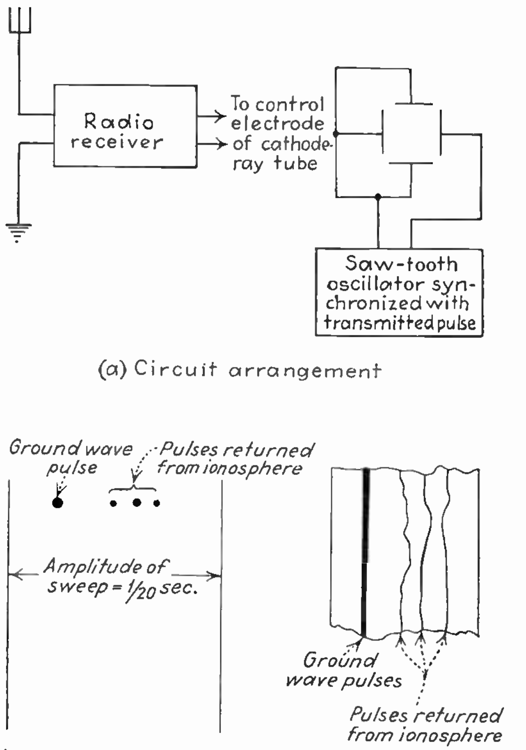



FIG. 182.—Arrangement of equipment for visually observing pulses received in ionosphere investigations, together with a typical received pattern.

Another method of observing Kennelly-Heaviside-layer behavior which is particularly adapted to the making of continuous records over long periods of time is shown in Fig. 183.[1] Here the pulses and the linear sweep circuit are synchronized with the power system as before, but the control electrode of the cathode-ray tube is biased beyond cut-off and so normally gives no spot. The output of the radio receiver is coupled to the control electrode in such a way that during the time a pulse is received the bias on the control electrode is reduced, thus permitting a momentary spot to exist. The resulting pattern is as shown in Fig. 183, consisting of a series of dots spaced along the time axis at distances from the ground-wave pulse corresponding to the delay times, and having an intensity which is roughly proportional

[1] See Lal C. Verman, S. T. Char, and Aijaz Mohammed, Continuous Recording of Retardation and Intensity of Echoes from the Ionosphere, *Proc. I.R.E.*, vol. 22, p. 906, July, 1934.

to the intensity of the received pulse. By photographing the pattern on a slowly moving film it is possible to obtain a continuous record of the various received pulses such as shown at Fig. 183*c*.




(*a*) Circuit arrangement

(*b*) Typical received pattern (*c*) Record on moving film

FIG. 183.—Arrangement of equipment that permits both visual observation and continuous photographic recording of pulses received in ionosphere investigations, together with typical received pattern.

Cathode-ray tubes are often used in measuring the characteristics of received waves. Thus Friis[1] used a cathode-ray tube to determine the direction of arrival of high-frequency signals with an arrangement of equipment such as shown in Fig. 184. Here two antennas with separate receivers and a common beating

[1] See H. T. Friis, Oscillographic Observations on the Direction of Propagation and Fading of Short Waves, *Proc. I.R.E.*, vol. 16, p. 658, May, 1928.

oscillator are required, with the receiver outputs applied to the cathode-ray tube as shown. Thus a wave arriving along the path *aa* combines with the local oscillator in the same phase in both antennas. In contrast to this a wave arriving along the path *bb* strikes one antenna before the other, whereas the local oscillations do not, thus causing the phase relations in the two antennas

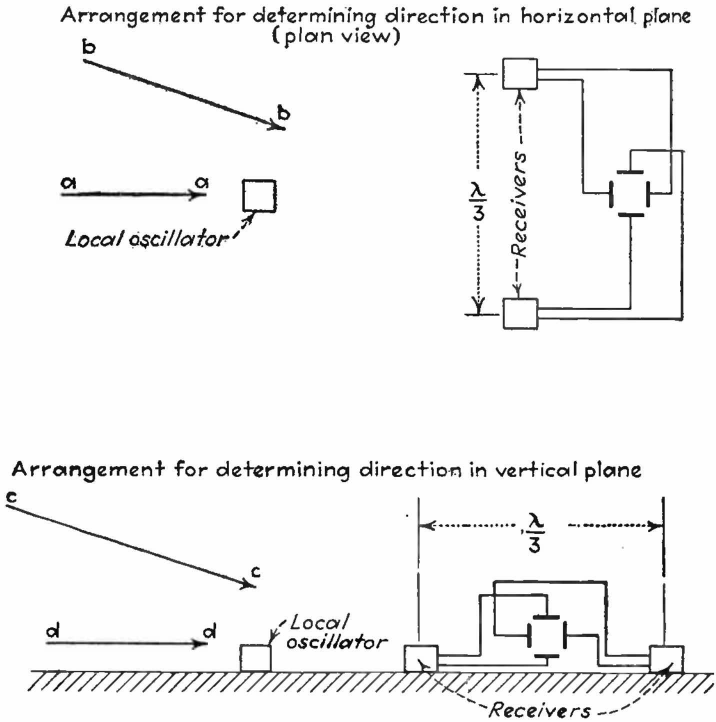



FIG. 184.—Method of using cathode-ray tube to determine direction of arrival of radio waves.

to differ. This phase difference is not affected by the beating operation and appears unaltered in the audio-frequency beat note that appears in the receiver output. Hence if a wave is arriving along the path *aa*, the pattern traced by the cathode-ray spot is a straight line since there is no phase shift, whereas waves arriving from other directions give an elliptical trace which can be used to calculate the angle of departure from *aa*.

Merritt and Bostwick[1] used a cathode-ray tube in a somewhat similar manner to investigate the structure of the received wave. In their equipment, one antenna was arranged to respond only to the horizontally polarized component of the sky wave while a second antenna system responded to the ground wave combined with the vertical polarized component of the sky wave. Separate

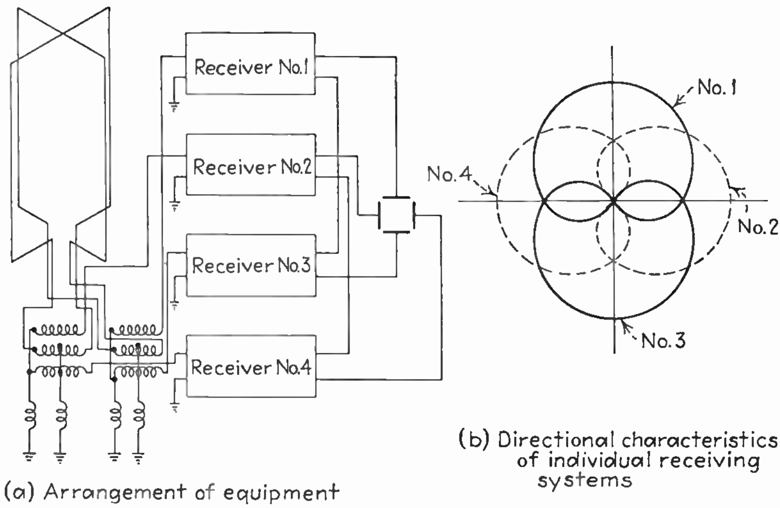



FIG. 185.—Circuit arrangement for giving instantaneous indication of the direction of arrival of radio signals and static pulses.

receivers were employed with each antenna system but both receivers made use of a common heterodyne oscillator. The receiver outputs were applied to the two pairs of deflecting plates. The patterns that result from such an arrangement are ellipses, and by an analysis of the antenna systems and the patterns it is possible to deduce information on the structure of the wave.

*Static Investigations with Cathode-ray Tubes.*—Cathode-ray tubes have been used to observe the wave form of individual static impulses.[2] This is accomplished by amplifying the static impulses with an aperiodic amplifier and applying the amplified output directly to one pair of deflectors. A suitable sweep

[1] E. Merritt and W. E. Bostwick, A Visual Method of Observing the Influence of Atmospheric Conditions on Radio Reception, *Proc. Nat. Acad. Sci.*, vol. 14, p. 884, November, 1928.

[2] Also see R. A. Watson-Watt, An Instantaneous Direct-reading Radiogoniometer, *Jour. I.E.E.*, vol. 64, p. 611, May, 1926.

voltage is applied to the other deflectors. If the individual static pulses do not arrive too rapidly and if the fluorescent screen of the cathode-ray tube is of a type having considerable persistence of illumination, the pattern produced by an individual pulse can be observed sufficiently well to permit sketching.

Observations on the direction of arrival of individual static pulses can be obtained with an arrangement of equipment as shown in Fig. 185. This consists essentially of two crossed-loop antennas arranged to give both loop and antenna action simultaneously, together with four separate radio receivers. The various receivers are connected to the antenna system in such a way that the response in each receiver is a cardioid, with the four cardioids oriented along four quadrants as shown in Fig. 185*b*. The receiver outputs are connected in pairs to the deflecting plates as shown with the polarities such that the individual receivers composing the pair deflect the spot in opposite directions. By making the directional characteristics true cardioids and adjusting the four receivers for equal sensitivity, a static impulse will cause the spot of the cathode-ray tube to dart out from the origin in the exact direction from which the static came.

# CHAPTER XIV

## MISCELLANEOUS

**80. Measurements on Power-supply Systems.**—The output voltage of a rectifier-filter system, and the variation of output voltage with load, can be readily determined with direct-current instruments. The results obtained will depend to some extent upon the wave form and magnitude of the supply voltage so that for accurate comparative results care must be taken to insure constant line-voltage conditions.

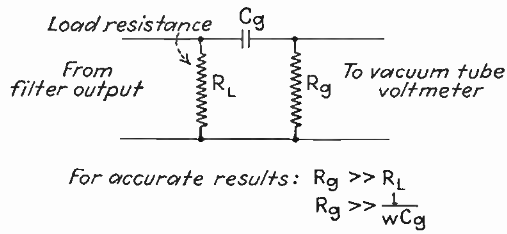



FIG. 186.—Arrangement for measuring the hum voltage developed across the output of a rectifier-filter system.

The hum voltage developed across the output of the filter system can be measured with a vacuum-tube voltmeter in a circuit such as illustrated in Fig. 186. The resistance-condenser combination blocks direct-current potentials from the measuring equipment and yet applies to the measuring equipment the full alternating hum voltage existing across the load, provided the reactance of the blocking condenser $C_g$ at the lowest hum frequency is small compared with the resistance $R_g$. By making the ratio $R_g/R_L$ large, the measuring circuit does not appreciably affect the load impedance.

The hum voltage appearing across the output of a rectifier-filter system is normally a substantially pure sine wave of the ripple frequency provided the rectifier is symmetrical. If it is necessary to analyze the output voltage exactly, this can be done by using one of the methods described in Chap. VI, employing an amplifier if necessary to obtain sufficient sensitivity. In carrying out such measurements, care must be taken to avoid altering the impedance relations existing across the load. Thus the harmonic-analyzer input should either offer the same impedance to currents of all frequencies, including direct currents, or should be so arranged that it consumes negligible energy. Results obtained from harmonic analysis of the hum voltage are chiefly of value in detecting the presence of unbalance in the rectifier, which is indicated by hum voltage of the power-supply frequency, and in detecting high harmonics of the ripple frequency which may be troublesome in special applications.

**81. Shielding.**—Practically perfect electrostatic shielding may be obtained by the use of a metal box. Copper, aluminum, brass, sheet iron, etc., are all satisfactory for the shield, and if the joints make perfect contact the degree of shielding obtained is very high. Electrostatic fields can also be shielded fairly well by using copper screen.

Magnetic flux can be controlled by shields of magnetic material or by non-magnetic conducting shields having low resistivity. Shields of magnetic material operate by short-circuiting the flux lines that tend to pass through the shield, and hence have an effectiveness that depends upon the thickness and upon the magnetic permeability at *low flux densities.* The degree of shielding obtained in this way by using ordinary magnetic materials is not very great because of the low permeability at low flux densities, and even when high-permeability material such as permalloy is employed it is difficult to obtain effective shielding against strong fields. Magnetic shields find their chief usefulness in protecting against unidirectional and low-frequency fields.

Conducting shields operate by taking advantage of the fact that, when magnetic flux cuts across a conductor, eddy currents are produced which oppose the penetration of the flux through the conductor. Thus by inclosing the space to be protected with a copper or aluminum box, alternating magnetic fields will be prevented from penetrating through the conducting walls of the box. To be effective, conducting shields must be of low-resistance material and must not be too thin. Copper is the most satisfactory material, with aluminum next, while brass, sheet iron, etc., are relatively unsatisfactory. Conducting shields are much more effective than magnetic shields at radio frequencies and at the higher audio frequencies, but are not effective with unidirectional or very low-frequency fields. The dividing line occurs somewhere in the lower audio range but depends somewhat upon circumstances.

A particularly difficult problem of shielding is encountered in attempting to protect low-level audio-frequency transformers against 60- and 120-cycle hum fields produced by power transformers and filter reactors. Inclosing the low-level transformer in a cast-iron case helps somewhat, but the low permeability of cast iron at low flux densities makes the shielding relatively imperfect even when the thickness is great. High-permeability

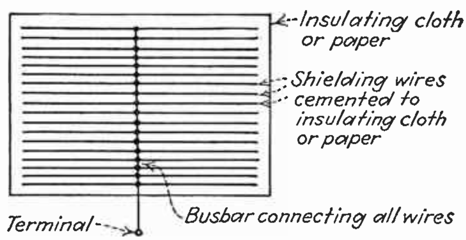



FIG. 187.—Electrostatic shield that does not affect magnetic flux.

shields (such as permalloy) are an improvement but are expensive and cannot be readily given the heat treatment required to develop their magnetic properties to the fullest extent. Conducting shields consisting of thick-walled copper boxes have been reported by several experimenters as more satisfactory than iron shields for 60- and 120-cycle magnetic fields.

When the shielding required must be complete, it is desirable to use several concentric shielding boxes of moderate thickness rather than one very thick box. When shielding against alternating magnetic and electrostatic fields, the shields should be either insulated from each other or connected together at only one point. A more detailed discussion of the problems involved in obtaining complete shielding of radio fields is given in Sec. 53.

There are circumstances when it is desirable to shield against electrostatic fields without interfering in any way with the magnetic fields which are present. This can be accomplished by the use of a conducting shield arranged so as to avoid closed loops around which magnetic flux can cause eddy currents to circulate. An example of such a shield is given in Fig. 187, where the shielding wires are connected together at only one point. Such a shield has no effect on magnetic flux, since conducting magnetic shields obtain their effectiveness from eddy currents and the arrangement of Fig. 187 provides no complete circuit around which eddy currents can flow. An electrostatic shield such as illustrated in Fig. 187 can be constructed by wrapping a single layer of varnished cambric or other insulator on a cylindrical form, and then winding a single, spaced layer of bare copper wire. A stiff busbar is laid across the wire, parallel to the axis of the cylinder, and soldered to each turn. The whole form is then painted with Duco or similar "dope" to bind the wire to the varnished cambric. After drying, the cambric with its attached wires is slipped off the form and cut parallel to the busbar.

**82. Construction of Transformers and Choke Coils.**—In laboratory work, special transformers and choke coils not commercially available are frequently required. Such transformers and choke coils can be constructed by using commercial laminations and winding the coils on a lathe. Laminations of material suitable for audio-frequency transformers, small power transformers, filter chokes, etc., are available already punched in *E-I* and other combinations, in a wide variety of dimensions. These laminations can be bought in small quantities at reasonable prices, and it is desirable to keep on hand a small supply of several representative sizes. Laminations of nickel-iron alloys[1] having a permeability that reaches as high as 100,000 can also be obtained, and are desirable for special applications.

A satisfactory method of winding coils for such laminations makes use of the form illustrated in Fig. 188. The core of this form is of the same dimension as the iron core with which the coil is to be used, and one endpiece can be removed by screws. In winding a coil one first winds one or two spacing layers of No. 18 to No. 22 wire, upon the top of which the desired coil is wound, using a lathe. The feed may be either by hand or by screw-thread. If it is desired that the winding be kept in regular layers, it is necessary to place gummed paper over each layer unless the wire is large. After winding, the entire coil form is

[1] A number of alloys of this general group have been developed and are commercialized under such names as permalloy, hypernik, A metal, Allegheny electric metal.

dipped in molten beeswax or in a mixture of beeswax and paraffin, and kept immersed until all bubbling ceases. This impregnation makes the coil self-supporting and moisture proof. An endpiece of the form is now removed, and the spacing layer removed by heating the core of the form with a soldering iron to soften the wax that holds this layer in place, and pulling out the spacing layers. This frees the coil, which can now be provided with permanent leads and taped.

Most wire companies are equipped to wind coils to specification. The charge for doing this normally consists of a small sum for setting up the winding machine, plus a charge for each coil.

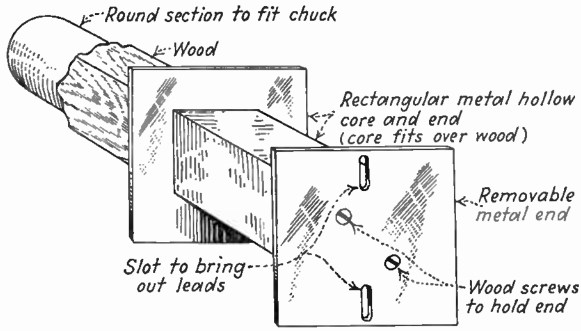



FIG. 188.—Form for winding coils for transformers and choke coils.

The charge per coil is little if any more than the cost of the same wire obtained through ordinary trade channels, and if a half dozen or more coils of the same size are desired it is usually more economical to purchase them by special order than it is to wind them by hand.

*Shielded and Balanced Transformers.*—Circumstances sometimes arise, particularly in connection with measuring equipment, where it is necessary to have an electrostatic shield between the primary and secondary of an audio-frequency transformer. Such shielding can be obtained by wrapping a one-turn conducting sheet of thin material between the windings as illustrated in Fig. 189, care being taken to insulate the overlapping point so that the shield does not act as a short-circuited turn. If desired, one or both windings can be completely inclosed in a conducting shield, care being taken to provide a break in the shield in a direction at right angles to the turns to prevent the shield from serving as a short-circuited turn that would introduce magnetic as well as electrostatic shielding. In some cases a double set of shields, such as illustrated in Fig. 54, is desirable. Electrostatic shields for transformers can be made of foil, shim brass, thin copper sheet, etc.

In certain types of transformers it is necessary that at least one winding be so arranged that the capacities to ground are balanced with respect to the terminals. This can be accomplished by the use of an electrostatic shield, and by dividing the winding into two identical halves which are exactly symmetrical with respect to the shield, as shown in Fig. 189. In order to

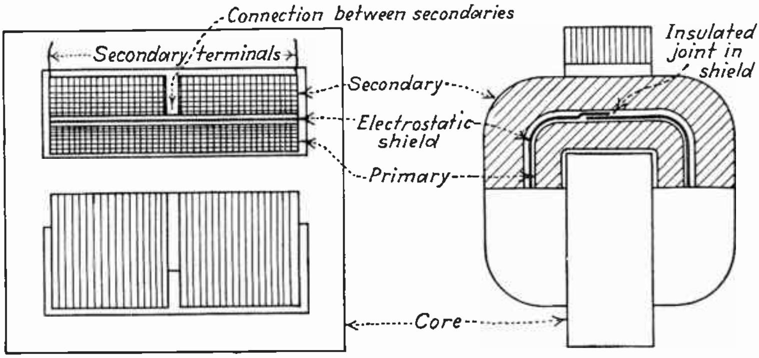



FIG. 189.—Transformer having a balanced secondary winding and an electrostatic shield between primary and secondary.

preserve the symmetry with respect to ground, the coils must be so assembled that the connections between halves are as illustrated in Fig. 189 when series aiding. The capacities between winding and shield can be kept as small as desired by using adequate clearances.

**83. Constructional Suggestions.**—In the building of laboratory equipment the constructional details are fully as important as the electrical design. The parts must be arranged in relation to each other in such a way as to minimize stray couplings, keep the length of critical leads as short as possible, etc., following the usual practices employed in the construction of radio receivers and high-gain speech amplifiers. It is also highly important that the method of mounting provide a maximum of accessibility to facilitate the testing, adjusting, servicing, and altering of the apparatus. It is particularly desirable to use a method of construction that makes it possible to have full access to the equipment while it is in actual operation. In general, compactness is of less importance than accessibility in laboratory apparatus. The use of ample space minimizes trouble from stray couplings, makes the parts more accessible, and provides room for minor alterations or additions. It is only when laboratory equipment is being built in quantities, for sale, that the question of maximum compactness becomes important.

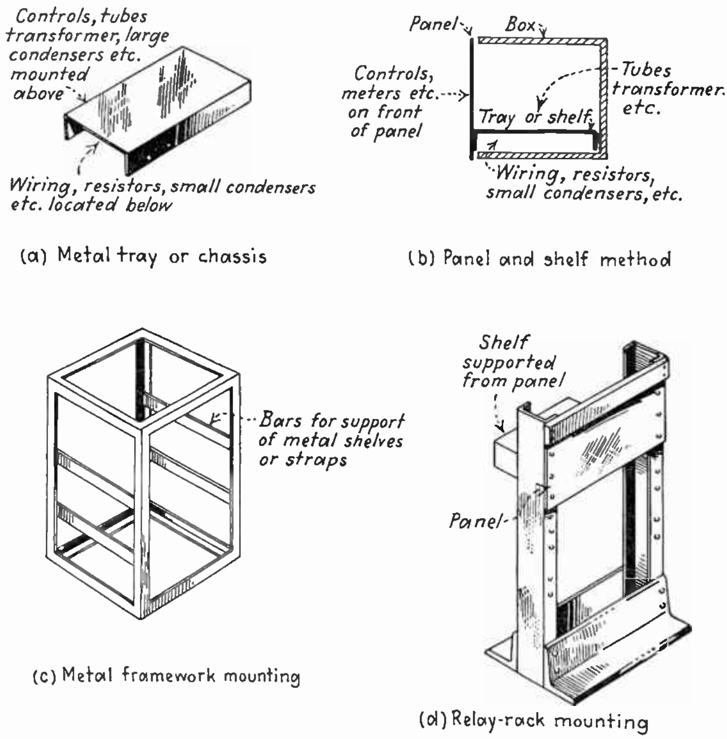



FIG. 190.—Standard methods of mounting laboratory apparatus.

The method of mounting should normally be one of the standard systems illustrated in Fig. 190. The metal-tray or chassis arrangement is used in broadcast receivers and has the advantage of accessibility, but does not lend itself to rugged mounting in a cabinet.

A modification of the metal-tray method is shown in Fig. 190*b*. Here the tray is mounted solidly to a metal or bakelite panel and this whole unit is then mounted in a metal or wood box by screws passing through the panel. The controls, meters, etc., are mounted on the panel while the remaining parts can be divided between the panel and the tray as desired. In this method of construction, the entire unit can be removed from its cabinet while operating by removing several screws.

In some cases it is possible to dispense entirely with the tray of Fig. 190*b* and mount everything upon the panel, with the tubes, controls, and meters in front of the panel and transformers, chokes, condensers, resistances, etc., behind. The panel is screwed to a wood or metal box which acts as a dust protection but which can be removed without disturbing the functioning of the equipment.

The mounting methods of Figs. 190*c* and 190*d* are used when the apparatus is either too large or too heavy to be handled by the simple methods already described, or when a number of closely associated units are involved. The framework illustrated in Fig. 190*c* is usually built up of angle iron or light channel pieces which are welded or bolted together. Equipment can be mounted on metal shelves and on strap-iron cross-pieces as is convenient. The front of the unit consists of a permanently mounted metal panel which carries the controls and meters. The sides, top, and back of the framework are covered with thin metal sheets that can be removed for testing or servicing the equipment. Where ventilation is important, the sides and back can be made of perforated metal. This general type of construction is commonly used on large portable or semiportable amplifiers and radio transmitters. High-power broadcast transmitters and code transmitters also use the same method of construction with modifications.

In the relay-rack method of construction illustrated in Fig. 190*d*, the various units are mounted upon panels or a combination of panels and metal trays, and these individual units are then mounted one above the other upon a relay rack. In this way any individual unit can be removed for testing, servicing, or rebuilding without altering the remaining units in any way. The connecting leads are cabled and carried up the side of the rack, usually inside the vertical structural-steel member, which is often provided with rings to support the wiring. The relay rack itself consists of two vertical structural-steel members, usually channel irons. For permanent insulations these vertical members can be fastened to the floor and ceiling, but for semi-portable arrangements such as illustrated in Fig. 190*d* two heavy angle irons are commonly used for the base with a piece of strap iron across the top. The weight and size of the individual members depend primarily upon the height of the rack. The width is usually made to fit a standard panel having a width of 19½ inches. There is also a standard method of drilling the channel pieces to accommodate the panels, which come in standard heights. The relay-rack method of mounting equipment is standard for telephone equipment, and for related apparatus such as broadcast control panels.

The metal trays and metal boxes often used in the construction of laboratory apparatus can be bent up from ordinary galvanized sheet iron. The thinnest material that will be satisfactory should be used since thin metal is by far the easiest to handle. Panels can also be made of heavy galvanized sheet metal, although when the panel is very large, or the parts to be mounted upon it are very heavy, black sheet iron 1/16 inch thick or heavier may be required. When light weight is important, boxes, panels, and even frameworks requiring angles and channels can be built up from aluminum or duralumin.

A satisfactory finish for metal panels and boxes can be obtained by using one of the baked lacquers that crystallizes or wrinkles on drying. These are usually sprayed on and can be baked under a gas flame in a sheet-iron oven. Some of the crystallizing lacquers can be painted on with a brush and air dried, but when so handled are not particularly durable.

Wooden boxes for mounting equipment can be constructed from oak and stained as desired. Where shielding is necessary, this can be provided by lining the inner side of the box with thin copper or galvanized iron.

Bakelite sheet as normally purchased is not particularly suitable for panels, etc., because its high gloss shows dirt and finger marks. Such panels should be finished by rubbing with steel wool to produce a grained surface, care being taken always to rub in the same direction, and then wiping with an oily rag.

Laboratory equipment should if possible be designed so that the panels carrying controls, meters, etc., are vertical. In this way, these surfaces, which are always difficult to clean, collect a minimum of dust.

Wiring in laboratory equipment should always be cabled as far as possible. In particular, all leads at ground potential to high-frequency currents, such as power-supply leads and other leads that have already been by-passed to ground, should be arranged in cables in order to simplify the wiring layout, reduce couplings, and improve the appearance. The cabling can be done by using ordinary stout string that has been boiled in beeswax. The stitch employed should be that illustrated in Fig. 191*a* rather than that of Fig. 191*b*, since the latter will unravel if the string breaks, whereas the former will not.

Labels always present a problem with homemade equipment. The best method is to engrave directly on the panels, or on

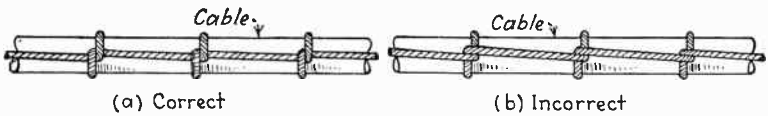



FIG. 191.—Lock stitch used in cabling wires.

bakelite strips that can be fastened to metal panels. This is both expensive and time consuming, however. After some experimenting the author has finally come to using labels printed with india ink, or typed with a fresh black ribbon, upon heavy bond paper. These labels are coated with clear Duco or similar lacquer to give a surface that can be easily cleaned, and are then trimmed to size and mounted on the panel with rubber cement. The Duco coat can be applied directly with a brush on drawing ink but must be carefully sprayed on typed labels to avoid smearing.

**84. Wave Filters.**—A wave filter is a combination of inductances and capacities which is so proportioned that, when this network is inserted between a generator and a load impedance, the presence of the network causes relatively little reduction in load power for a certain range of frequencies, while causing a very great reduction in load power at other frequencies. The outstanding characteristics of a properly designed wave filter are the relatively little effect that the presence of the filter has on those frequencies that it transmits effectively, the very great reduction in output or attenuation that the filter produces for other frequencies, and the very sharp transition from the transmitting to the attenuating condition.

The most important types of filters are low-pass, high-pass, and band-pass filters. A low-pass filter transmits all frequencies equally well from zero up to a certain limiting or cut-off frequency, and attenuates all higher frequencies. A high-pass filter transmits all frequencies from infinity down to a certain limiting or cut-off frequency, and attenuates all lower frequencies. A band-pass filter transmits nearly equally well all frequencies lying between two critical or cut-off frequencies, but attenuates frequencies both higher and lower than this band.

Ordinary filters normally contain two distinct parts. *First,* there are the terminating sections located at the input and output ends and designed to prevent the filter from introducing resonances. *Second,* there are intermediate sections placed between the end sections, and designed to build the attenuation characteristic up to the desired value. The filter may be built up on the basis of either $\pi$ or T intermediate sections, and in either case must be so proportioned that all parts cooperate with each other and the load resistance to give the desired performance. The design procedure required to accomplish this is discussed below, together with the necessary formulas.[1]

*Design Formula for Low-pass Filters.*—The formulas necessary for designing low-pass filters are given in Table VII*A* where the intermediate sections are of the T type, and in Table VII*B* where they are of the $\pi$ type. In each case the terminal half sections are designed according to the formulas given, using a value of design constant $m$ which is approximately 0.6. Half sections designed in this way prevent appreciable resonance in the pass band of frequencies when the proper load resistance is used. These terminal half sections have an attenuation

[1] It is beyond the scope of this book to develop the complete theory of the wave filter. All that can be done here is to give the formulas necessary to designing filters, and to show how they should be used. For further information concerning the theory and the calculation of performance, the reader should consult one or more of the following references: K. S. Johnson, "Transmission Circuits for Telephonic Communication," D. Van Nostrand Company, 1925; T. F. Shea, "Transmission Networks and Wave Filters," D. Van Nostrand Company, 1929; W. S. Franklin and F. E. Terman, "Transmission Line Theory," Franklin & Charles, 1926; O. J. Zobel, Theory and Design of Uniform and Composite Wave Filters, *Bell System Tech. Jour.*, vol. 2, p. 1, January, 1923; O. J. Zobel, Transmission Characteristics of Electric Wave Filters, *Bell System Tech. Jour.*, vol. 3, p. 567, October, 1924.

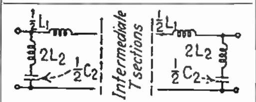
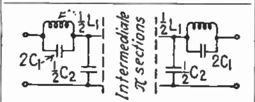
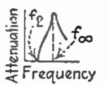
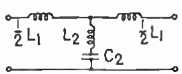
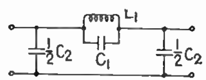
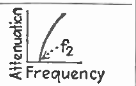
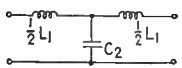
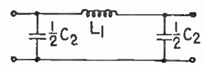
## Table VII.—Design of Low-pass Sections

### Fundamental Relations

$R$ = load resistance $\quad$ $f_2$ = cut-off frequency (highest frequency transmitted) $\quad$ $f_\infty$ = a frequency of very high attenuation

$$L_k = \frac{R}{\pi f_2} \qquad C_k = \frac{1}{\pi f_2 R} \qquad m = \sqrt{1 - \left(\frac{f_2}{f_\infty}\right)^2}$$

### Design of Sections

| Type | Attenuation characteristic | *A*. Filters having T intermediate sections | | *B*. Filters having $\pi$ intermediate sections | |
|---|---|---|---|---|---|
| | | Configuration | Formulas | Configuration | Formulas |
| End ($m$ of approximately 0.6) |   |   | $L_1 = mL_k$<br>$L_2 = \dfrac{1 - m^2}{4m}L_k$<br>$C_2 = mC_k$ |   | $L_1 = mL_k$<br>$C_1 = \dfrac{1 - m^2}{4m}C_k$<br>$C_2 = mC_k$ |
| I |   |   | $L_1 = mL_k$<br>$L_2 = \dfrac{1 - m^2}{4m}L_k$<br>$C_2 = mC_k$ |   | $L_1 = mL_k$<br>$C_1 = \dfrac{1 - m^2}{4m}C_k$<br>$C_2 = mC_k$ |
| II ($f_\infty = \infty$) |   |   | $L_1 = L_k$<br>$C_2 = C_k$ |   | $L_1 = L_k$<br>$C_2 = C_k$ |

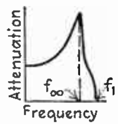
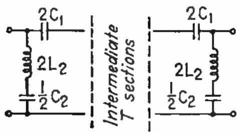
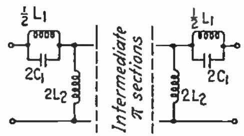
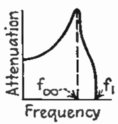
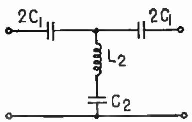
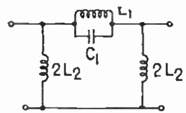
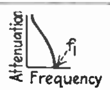
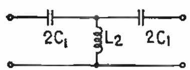
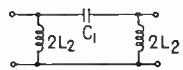
## Table VIII.—Design of High-pass Sections

### Fundamental Relations

$R_1$ = load resistance $\quad$ $f_1$ = cut-off frequency (lowest frequency transmitted) $\quad$ $f_\infty$ = a frequency of very high attenuation

$$L_k = \frac{R}{4\pi f_1} \qquad C_k = \frac{1}{4\pi f_1 R} \qquad m = \sqrt{1 - \left(\frac{f_\infty}{f_1}\right)^2}$$

### Design of Sections

| Type | Attenuation characteristic | *A.* Filters having T intermediate sections | | *B.* Filters having $\pi$ intermediate sections | |
|---|---|---|---|---|---|
| | | Configuration | Formulas | Configuration | Formulas |
| End ($m$ of approximately 0.6) |   |   | $C_1 = \frac{C_k}{m}$<br>$C_2 = \frac{4m}{1-m^2}C_k$<br>$L_2 = \frac{L_k}{m}$ |   | $L_1 = \frac{4m}{1-m^2}L_k$<br>$C_1 = \frac{C_k}{m}$<br>$L_2 = \frac{L_k}{m}$ |
| I |   |   | $C_1 = \frac{C_k}{m}$<br>$C_2 = \frac{4m}{1-m^2}C_k$<br>$L_2 = \frac{L_k}{m}$ |   | $L_1 = \frac{4m}{1-m^2}L_k$<br>$C_1 = \frac{C_k}{m}$<br>$L_2 = \frac{L_k}{m}$ |
| II $f_\infty = 0$ |   |   | $C_1 = C_k$<br>$L_2 = L_k$ |   | $C_1 = C_k$<br>$L_2 = L_k$ |

characteristic approximately that as shown, with the frequency of high attenuation $f_\infty$ fixed in relation to the cut-off frequency $f_2$ by the fact that $m$ is approximately 0.6.

The intermediate sections placed between the terminal half sections are for the purpose of supplementing the attenuation of the end sections so that a higher and more uniform attenuation may be obtained. These sections may be of either the Type I or Type II kind, depending upon the attenuation characteristic desired. In the Type I section the frequency at which high attenuation takes place can be controlled by varying the design constant $m$. In this way, by using several sections with different values of $m$, it is possible to make the attenuation characteristics of the intermediate sections supplement each other and also supplement the end sections. It will be noted that the Type II section is a special case for which $m = 1.0$, which makes the frequency of high attenuation equal to infinity.

*Design Formula for High-pass Filters.*—Design formulas for high-pass filters using the T and $\pi$ arrangements of intermediate sections are given in Tables VIII*A* and VIII*B*, respectively. The terminal half sections are designed according to the formulas given, with the design constant $m$ equal to approximately 0.6. These half sections, as in the case of the low-pass filter, are for the purpose of preventing resonances in the pass band when the filter is used with the proper load resistance, and their attenuation characteristic is fixed by the fact that the design constant $m$ must be approximately 0.6.

The intermediate sections which can be placed between the terminal half sections are for the purpose of supplementing the attenuation of the end sections to give higher and more uniform attenuation. These sections may be of either the Type I or Type II kind depending upon the attenuation characteristics desired. In the Type I section it will be noted that the frequency at which the high attenuation takes place is an independent variable and can be assigned any arbitrary value as desired. In this way it is possible to design one or more intermediate sections which have attenuation characteristics that differ from each other, and from the end sections, in such a way as to supplement each other. It will be noted that the Type II section is a Type I section where the frequency of high attenuation has been assigned the value zero.

# TABLE IX.—DESIGN OF BAND-PASS SECTIONS

## Fundamental Relations

$R$ = load resistance

$f_1$ = lower frequency limit of pass band

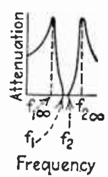
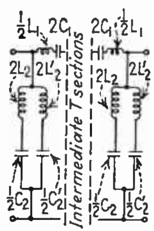
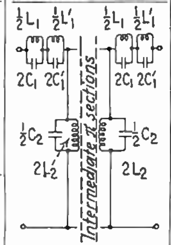
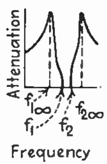
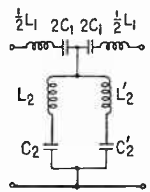
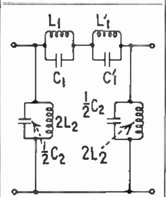
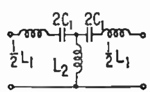
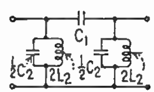
$f_2$ = higher frequency limit of pass band

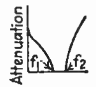
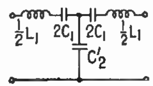
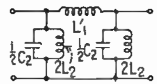
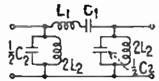
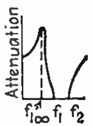
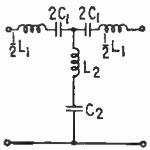
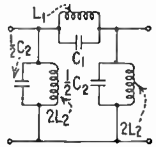
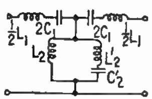
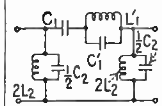
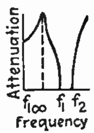
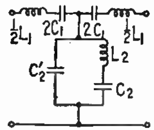
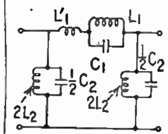
$f_{1\infty}$ = a frequency of very high attenuation in low-frequency attenuating band

$f_{2\infty}$ = a frequency of very high attenuation in high-frequency attenuating band

$$L_{1k} = \frac{R}{\pi(f_2 - f_1)} \qquad L_{2k} = \frac{(f_2 - f_1)R}{4\pi f_1 f_2} \qquad C_{1k} = \frac{f_2 - f_1}{4\pi f_1 f_2 R} \qquad C_{2k} = \frac{1}{\pi(f_2 - f_1)R}$$

## Design of Sections

| Type | Attenuation characteristic | *A.* Filters having T intermediate sections: Configuration | *A.* Filters having T intermediate sections: Formulas | *B.* Filters having $\pi$ intermediate sections: Configuration | *B.* Filters having $\pi$ intermediate sections: Formulas | Notation for both T and $\pi$ sections |
|---|---|---|---|---|---|---|
| End ($m_1 = m_2$ = approximately 0.6) |  |  | $L_1 = m_1 L_{1k}$<br>$L_2 = aL_{1k}$<br>$L_2' = cL_{1k}$ $C_1 = \frac{C_{1k}}{m_2}$<br>$C_2 = \frac{C_{1k}}{b}$ $C_2' = \frac{C_{1k}}{d}$ |  | $L_1 = \frac{L_{2k}}{b}$<br>$L_1' = \frac{L_{2k}}{d}$<br>$L_2 = \frac{L_{2k}}{m_2}$ $C_1 = aC_{2k}$<br>$C_1' = cC_{2k}$<br>$C_2 = m_1 C_{2k}$ | $g = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f_1^2}\right)\left(1 - \frac{f_{1\infty}^2}{f_2^2}\right)}$<br>$h = \sqrt{\left(1 - \frac{f_1^2}{f_{2\infty}^2}\right)\left(1 - \frac{f_2^2}{f_{2\infty}^2}\right)}$<br>$m_1 = \frac{\frac{f_1 f_2}{f_{2\infty}^2}g + h}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}$ $m_2 = \frac{g + \frac{f_{1\infty}^2}{f_1 f_2}h}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}$<br>$a = \frac{(1-m_1^2)f_{2\infty}^2}{4gf_1f_2}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{(1-m_2^2)f_1f_2}{4gf_{1\infty}^2}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$<br>$b = \frac{(1-m_2^2)}{4g}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ $c = \frac{(1-m_1^2)}{4h}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$<br>$d = \frac{(1-m_1^2)f_{2\infty}^2}{4hf_1f_2}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{(1-m_2^2)f_1f_2}{4hf_{1\infty}^2}\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$<br>when $(m_1 = m_2)$,<br>$g = h, a = d, b = c, f_{1\infty} = \frac{f_1 f_2}{f_{2\infty}}, m_1 = m_2 = \frac{h}{1 - \frac{f_1 f_2}{f_{2\infty}^2}}$<br>and<br>$f_{2\infty}^2 = \frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1-m^2)} + \left[\left(\frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1-m^2)}\right)^2 - f_1^2 f_2^2\right]^{1/2}$<br>These formulas apply for both end and Type I sections |
| I |  |  | $L_1 = m_1 L_{1k}$<br>$L_2 = aL_{1k}$<br>$L_2' = cL_{1k}$ $C_1 = \frac{C_{1k}}{m_2}$<br>$C_2 = \frac{C_{1k}}{b}$ $C_2' = \frac{C_{1k}}{d}$ |  | $L_1 = \frac{L_{2k}}{b}$ $L_1' = \frac{L_{2k}}{d}$<br>$L_2 = \frac{L_{2k}}{m_2}$ $C_1 = aC_{2k}$<br>$C_1' = cC_{2k}$<br>$C_2 = m_1 C_{2k}$ | |
| II $f_{1\infty} = 0$ $f_{2\infty} = f_2$ |  |  | $L_1 = \frac{f_1 R}{\pi f_2(f_2 - f_1)}$<br>$L_2 = \frac{(f_1 + f_2)R}{4\pi f_1 f_2}$<br>$C_1 = C_{1k}$ |  | $C_1 = \frac{f_1 + f_2}{4\pi f_1 f_2 R}$<br>$C_2 = \frac{f_1}{\pi f_2(f_2 - f_1)R}$<br>$L_2 = L_{2k}$ | |

| | | | | | | |
|---|---|---|---|---|---|---|
| III $f_{1\infty} = f_1$ $f_{2\infty} = \infty$ |  |  | $L_1 = L_{1k}$ $C_2' = \frac{1}{\pi(f_1 + f_2)R}$ $C_1 = \frac{f_2 - f_1}{4\pi f_1^2 R}$ |  | $L_1' = \frac{R}{\pi(f_1 + f_2)}$ $L_2 = \frac{(f_2 - f_1)R}{4\pi f_1^2}$ $C_2 = C_{2k}$ | |
| IV $f_{1\infty} = 0$ $f_{2\infty} = \infty$ |  |  | $L_1 = L_{1k}$ $L_2 = L_{2k}$ $C_1 = C_{1k}$ $C_2 = C_{2k}$ |  | $L_1 = L_{1k}$ $L_2 = L_{2k}$ $C_1 = C_{1k}$ $C_2 = C_{2k}$ | |
| V $f_{2\infty} = f_2$ |  |  | $L_1 = m_1 L_{1k}$ $L_2 = \frac{(1 - m_1^2)}{4m_1} L_{1k}$ $C_1 = \frac{C_{1k}}{m_2}$ $C_2 = \frac{4m_2}{1 - m_2^2} C_{1k}$ See notation for $m_1$ and $m_2$ |  | $L_1 = \frac{4m_2}{1 - m_2^2} L_{2k}$ $L_2 = \frac{L_{2k}}{m_2}$ $C_1 = \frac{(1 - m_1^2)}{4m_1} C_{2k}$ $C_2 = m_1 C_{2k}$ See notation for $m_1$ and $m_2$ | $m_1 = \frac{f_1}{f_2} m_2$ $m_2 = \sqrt{\frac{1 - \frac{f_{1\infty}^2}{f_1^2}}{1 - \frac{f_{1\infty}^2}{f_2^2}}}$ |
| VI $f_{1\infty} = f_1$ |  | Same circuit as above for Type V | Same formulas as above for Type V. See notation for $m_1$ and $m_2$ | Same circuit as above for Type V | Same formulas as above for Type V See notation for $m_1$ and $m_2$ | $m_1 = \sqrt{\frac{1 - \frac{f_2^2}{f_{2\infty}^2}}{1 - \frac{f_1^2}{f_{2\infty}^2}}}$ $m_2 = \frac{f_1}{f_2} m_1$ |
| VII $f_{1\infty} = 0$ |  |  | $L_1 = m_1 L_{1k}$ $L_2 = a L_{1k}$ $L_2' = \frac{(1 - m_1^2)}{4h} L_{1k}$ $C_1 = C_{1k}$ $C_2' = \frac{h}{a} C_{1k}$ |  | $L_1' = \frac{h}{a} L_{2k}$ $L_2 = L_{2k}$ $C_1 = a C_{2k}$ $C_2 = m_1 C_{2k}$ $C_1' = \frac{(1 - m_1^2)}{4h} C_{2k}$ | $h = \sqrt{\left(1 - \frac{f_1^2}{f_{2\infty}^2}\right)\left(1 - \frac{f_2^2}{f_{2\infty}^2}\right)}$ $m_1 = \frac{f_1 f_2}{f_{2\infty}^2} + h$ $a = \frac{(1 - m_1^2) f_{2\infty}^2}{4 f_1 f_2}$ |
| VIII $f_{2\infty} = \infty$ |  |  | $L_1 = L_{1k}$ $L_2 = \frac{d}{g} L_{1k}$ $C_1 = \frac{C_{1k}}{m_2}$ $C_2 = \frac{4g}{1 - m_2^2} C_{1k}$ $C_2' = \frac{C_{1k}}{d}$ |  | $L_1 = \frac{4g}{1 - m_2^2} L_{2k}$ $L_2 = \frac{L_{2k}}{m_2}$ $L_1^1 = \frac{L_{2k}}{d}$ $C_1 = \frac{d}{g} C_{2k}$ $C_2 = C_{2k}$ | $g = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f_1^2}\right)\left(1 - \frac{f_{1\infty}^2}{f_2^2}\right)}$ $m_2 = g + \frac{f_{1\infty}^2}{f_1 f_2}$ $d = \frac{(1 - m_2^2) f_1 f_2}{4 f_{1\infty}^2}$ |

*Design Formula for Band-pass Filters.*—Formulas for the design of band-pass filters employing T and $\pi$ types of intermediate sections are given in Tables IX*A* and IX*B*, respectively. The terminal half sections of these filters must be designed as shown, with the design constants $m_1$ and $m_2$ both equal to approximately 0.6. Such half sections cause the filter to operate with negligible resonances in the pass band of frequencies. The attenuation characteristics of the terminal sections are fixed by the fact that these sections must have values of $m_1$ and $m_2$ very nearly 0.6.

A wide variety of sections are available for use as intermediate sections in a band-pass filter, as is apparent from Table IX. The Type I section represents the most general case and is characterized by having a frequency of high attenuation in each pass band. The location of these frequencies of high attenuation can be varied as desired, thus making it possible to construct intermediate sections which have unlike attenuation characteristics and also attenuation characteristics that differ from those of the terminal sections. The remaining types of band-pass sections represent special cases for which one or more of the reactance elements degenerate into either an open or a short circuit. This results in simplification of the network but also is accompanied by a less favorable and less flexible attenuating characteristic. It can, in fact, be shown that the more complicated sections are fundamentally equivalent to two of the simple types of sections connected in series.

*Design Procedure.*—The design of a filter to meet a given set of conditions falls into several well-defined steps, which are briefly covered by the following headings:

1. Determine the cut-off frequencies (*i.e.*, the frequencies that mark the edge of the pass band) and the load resistance that will meet the design requirements, and decide whether T or $\pi$ intermediate sections are to be used.

2. Design the terminating sections according to the proper table, using the load resistance and cut-off frequencies selected in Step 1.

3. Decide on the number of intermediate sections to be used. The more sections selected, the greater will be the attenuation of those frequencies which are to be attenuated. One intermediate section will give a moderately good filter and is sufficient

for many conditions. More than two intermediate sections are never required except in rare instances.

4. Select the frequencies at which the different intermediate sections will have their maximum attenuation, and then design these sections for the chosen load resistance and cut-off frequencies, using the formulas in the appropriate table.

The first problem in the design of filters is a choice between T and $\pi$ sections for the intermediate sections. This decision is based primarily upon considerations of convenience, for electrically the performance is substantially identical. It will be found, however, that at the higher audio frequencies and also at radio frequencies it is preferable to use $\pi$ sections because when this is done the design normally calls for capacities in shunt with the coils, which permits the distributed capacity of the coil to be used as part of the required shunting capacity. If sections used do not call for capacity in shunt with the coils, then under conditions where the distributed capacity of the coils is appreciable it is apparent that the filter as actually built can never be exactly what is called for by the design tables.

The location of the frequencies for which the intermediate sections have high attenuation must be carefully selected, for upon this choice rest the attenuating properties of the filter. These frequencies of high attenuation should in general be different for the different sections and should be so staggered throughout the attenuating bands that every frequency suffers considerable attenuation by at least one section. By proper judgment it is possible so to arrange matters that when the terminating sections are augmented by only one or two intermediate sections there is a uniformly high attenuation over the entire attenuating bands, with no great surplus of attenuation at some frequencies and no serious lack at others. The terminal sections determine one frequency of high attenuation in each attenuating band, and so should be designed before the intermediate sections in order that the attenuation of these central sections may best supplement the end sections.

It is often desirable that one of the intermediate sections be so designed as to have a frequency of high attenuation very close to the cut-off frequency in order that the attenuation may rise sharply at the border of the pass band. While it is theoretically possible in the ideal case with no-loss reactances to make this frequency as close to the cut-off frequency as desired, practical considerations require that all frequencies of high attenuation differ by at least 2 to 5 per cent from the cut-off frequency. There are two reasons for this. In the first place, as the frequency of high attenuation is placed closer and closer to the cut-off frequency, the attenuating characteristic becomes narrower, as shown in Fig. 192, with the result that a section designed to attenuate frequencies extremely close to the cut-off frequency is of very little use in attenuating frequencies appreciably different from cut-off. In the second place, the reactance

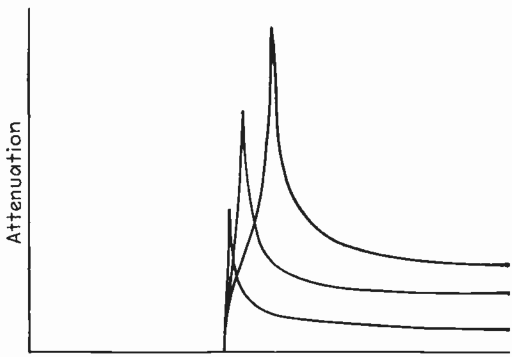



FIG. 192.—Attenuation characteristics obtained in a low-pass filter section as the frequency of high attenuation is placed closer and closer to cut-off.

arms required assume impracticable proportions if the frequency of high attenuation is placed very close to cut-off, and if these impractical circuit elements are built it will be found that the losses prevent the high attenuation from being realized. This is illustrated in Fig. 192.

In order that the full possibilities of a wave filter may be realized, it is necessary that the resistance of the reactance elements be reasonably low. The coils are the limiting factor in this regard and should if possible have a ratio of reactance to resistance that exceeds 15 and is preferably of the order of 25 to 100. Coils with ordinary laminated cores are entirely satisfactory at power and lower audio frequencies, and in fact give fair results throughout the voice range. Permalloy- and iron-dust rings such as used for telephone loading coils are particularly good for the audio-frequency and carrier-frequency ranges, and will normally give a ratio of reactance to resistance that is at least 50.

The principal problem involved with the filter condensers is in obtaining the exact capacity required. At audio frequencies this is most readily accomplished by the use of sectionalized telephone condensers, such as the Western Electric 57-AK and 134-A types, supplemented by large fixed condensers if this is necessary. At radio frequencies fixed mica condensers supplemented by adjustable trimming condensers can be employed.

In adjusting the inductances and capacities that enter into the filter it is desirable that the individual inductance and capacity values be as nearly correct as possible, but it is even more essential that the various resonance frequencies be correct than that the proper proportion of inductance and capacity be present. Thus in a filter section such as shown in Table IX*A*, the coils and condensers appear in pairs which form series-resonant circuits, and it is highly important that when the filter is built these resonant frequencies be of the proper value even though the exact design proportion of inductance and capacity may not be realized. Likewise in a section such as illustrated in Table IX*B*, the coils and condensers appear in pairs which form parallel resonant circuits, and it is highly important that the resonant frequency of each of these pairs have exactly the proper value when the filter is built.

*Examples.*—As an illustration of how the design procedure is carried out, and what type of performance may be expected, several typical examples will now be considered.

Consider first a low-pass filter which is to operate with a load resistance of 700 ohms, and to have a cut-off frequency of 1213 cycles. A suitable design employing T sections is shown in Fig. 193*a* and an equivalent design employing $\pi$ intermediate sections at Fig. 193*b*. The terminal half sections have been given a value $m = 0.6$, and as a result have their frequency of high attenuation at 1516 cycles. One intermediate section is designed so that the frequency of high attenuation is 1322 cycles, thereby giving a rapid rise of attenuation near cut-off. The second intermediate section is designed so that the frequency of high attenuation is at infinity, thereby insuring good attenuation for frequencies which differ considerably from cut-off. The resulting inductance and capacity values are as shown. In constructing the filter it will be noted that various inductances and also various capacities may be combined to reduce the

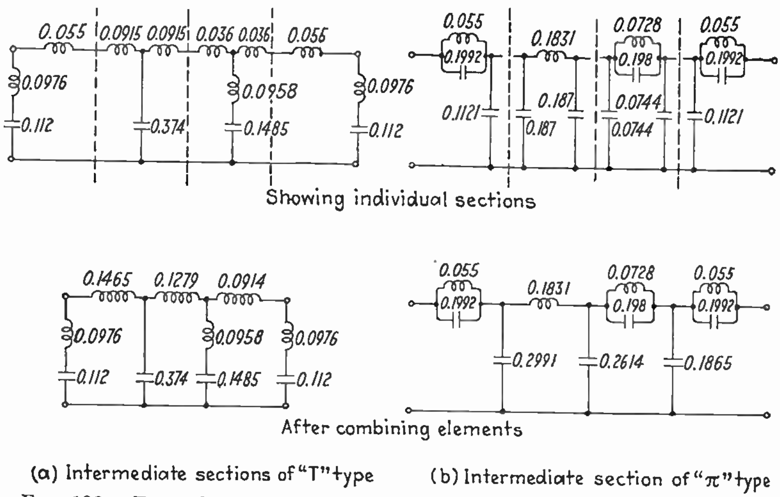



FIG. 193.—Example of a low-pass filter having a cut-off frequency of 1213 cycles, two intermediate sections, and proportioned to operate with a load resistance of 700 ohms.

number of coils and condensers required. This causes the identity of the individual sections to be lost but does not alter the performance in any way as far as the terminals are concerned.

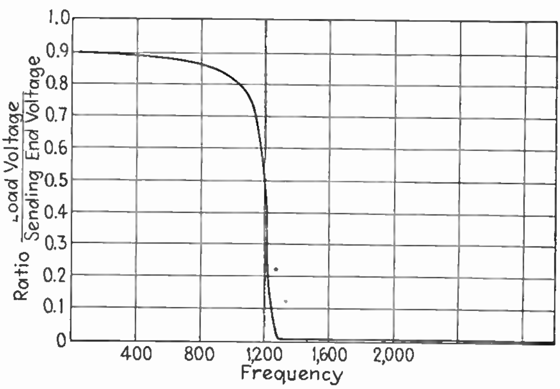



FIG. 194.—Ratio of voltage across load to voltage across sending end as measured experimentally for the filter of Fig. 193b.

The filter of Fig. 193b has been actually built using air-cored coils with rather high losses, and was found to have a transmission characteristic as shown in Fig. 194. The transmission is seen to cease very abruptly in the vicinity of the cut-off frequency and the output voltage is uniformly less than one-hundredth of the input voltage for all frequencies that differ from cut-off by at least 70 cycles.

A high-pass filter designed to operate with a load resistance of 700 ohms and have a cut-off frequency of 1000 cycles is illustrated in Figs. 195$a$ and 195$b$. The terminal half sections in each case are designed for $m = 0.6$ and so have a frequency

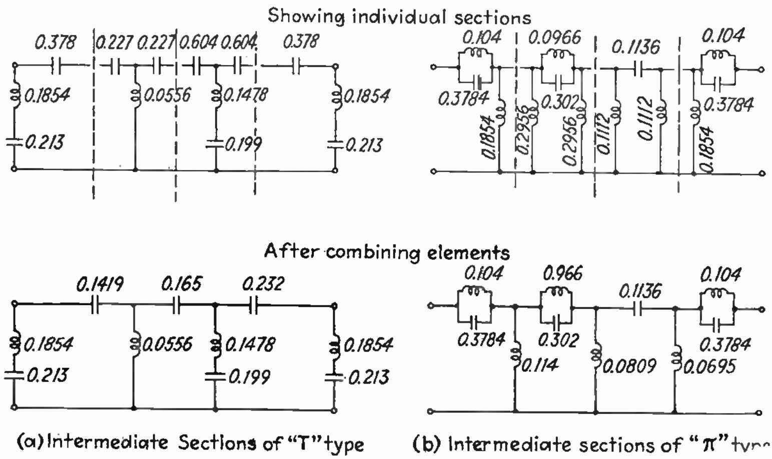



FIG. 195.—Example of a high-pass filter having a cut-off frequency of 1000 cycles, two intermediate sections, and proportioned to operate with a load resistance of 700 ohms.

of high attenuation at 800 cycles. Two intermediate sections are employed, one having a frequency of high attenuation at 925 cycles in order to insure a rapid rise of attenuation near cut-off, and one having its frequency of high attenuation at zero frequency in order to insure that frequencies much lower than cut-off will be adequately attenuated. The resulting sections have capacity and inductance values as shown, and can be combined as indicated. The filter composed of T intermediate sections was actually built using iron-dust cored coils, and was found to have the transmission characteristic shown in Fig. 196. The transmission is seen to cease abruptly in the vicinity of the cut-off frequency, and the output voltage is uniformly less than one-hundredth of the input voltage for all frequencies below 950 cycles.

A band-pass filter designed to operate with a load resistance of 700 ohms, and to transmit the band of frequencies lying

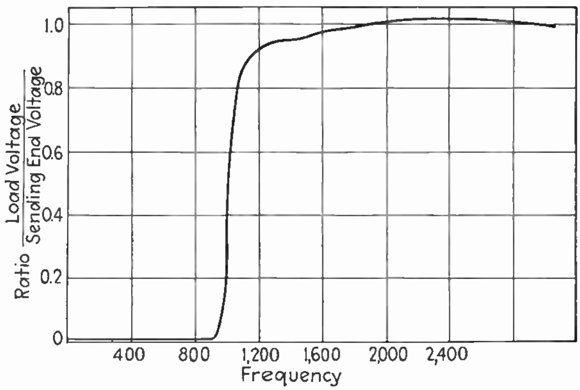



FIG. 196.—Ratio of voltage across load to voltage across sending end as determined experimentally for the filter of Fig. 195*a*.

between 500 and 2000 cycles, is shown in Fig. 197. The terminal half sections are designed for $m_1 = m_2 = 0.6$. This makes the frequencies of high attenuation 433 and 2310 cycles. It will be noted that these are sufficiently close to cut-off to provide

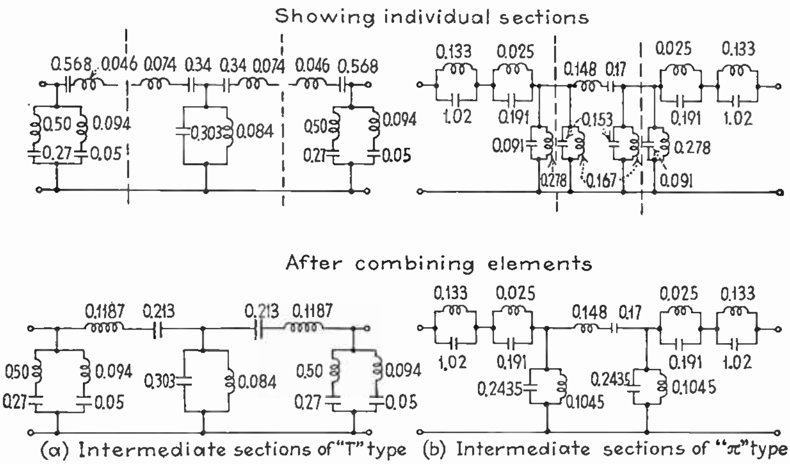



FIG. 197.—Example of a band-pass filter having cut-off frequencies of 500 and 2000 cycles, one intermediate section, and proportioned to operate with a load resistance of 700 ohms.

rapid rise of the filter attenuation near cut-off. The intermediate portion of the filter hence needs attenuate only those frequencies differing considerably from cut-off. The filters shown employ a single intermediate section designed so that the frequencies of high attenuation are zero and infinity. The resulting capacity and inductance elements are as shown in Fig. 197, and the transmission characteristic actually obtained from this filter when built with a $\pi$ intermediate section and using permalloy-dust cored coils is as shown in Fig. 198.

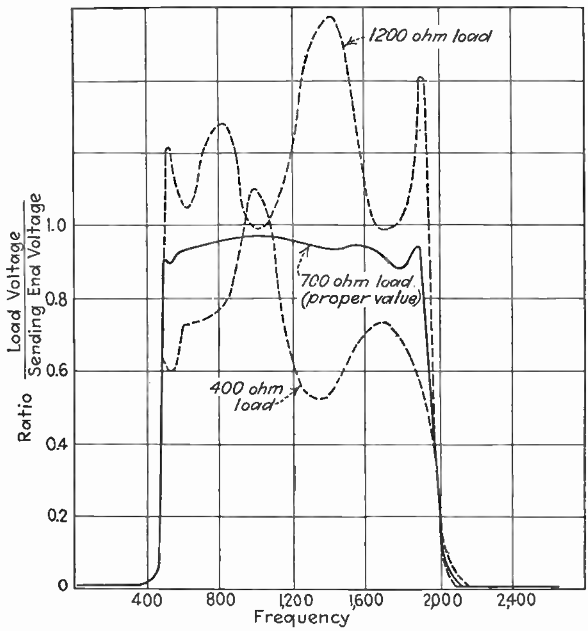



FIG. 198.—Ratio of voltage across load to voltage across sending end as determined experimentally for the filter of Fig. 197*b*, showing the resonances that result in the pass band when the improper load resistance is used.

*Miscellaneous Comments.*—A wave filter is essentially a carefully balanced combination of reactance elements so arranged that over a limited range of frequencies the resonance effects balance out. As a result it is essential that the design requirements be realized in actual operation if the full possibilities are to be realized. Thus if the incorrect load resistance is used, resonances will be introduced in the pass band of frequencies as illustrated in Fig. 198, and a similar result will occur if even a single element of the filter is appreciably out of adjustment. It is particularly important that terminal sections be employed, since if a filter is built up using only intermediate sections resonances similar to those shown in Fig. 198 will always occur in the pass band.

When a filter is properly designed and operated and employs coils with very low losses, the characteristics attained are remarkably close to perfection, as is shown by the performance of the filters in Figs. 194, 196, and 198. Under such circumstances the principal effect which the presence of the filter has upon the frequencies that are transmitted is to shift their phase somewhat.

The presence of losses in the filter reactances introduces effects which are qualitatively easily predicted. Losses cause some attenuation even for those frequencies which should be transmitted with a minimum of attenuation, and losses also tend to round off the abrupt transition from pass to attenuating conditions that would otherwise exist at the cut-off frequencies. The only important effect of the losses on the attenuating characteristics is to reduce the amount of attenuation obtainable at the frequencies of high attenuation. If the ratio of reactance to resistance of the various elements is at least 15 and preferably in excess of 25, the losses will exert only a modifying influence and will not alter the performance in essential respects.

# APPENDIX I

## LABORATORY EXPERIMENTS

### PRELIMINARY NOTE

These experiments are for the purpose of assisting those planning laboratory work dealing with vacuum-tube and radio phenomena. Since no two laboratories can be expected to possess the same apparatus, these experiments are outlined only in skeleton form, and the final details will have to be worked out in relation to each individual laboratory. No effort has been made to arrange the experiments so that they represent approximately equal amounts of laboratory work because of the fact that the time required will often depend greatly upon the equipment available. The experiments as outlined are in most cases rather long as a result of the fact that an effort has been made to indicate all of the things that can be done without undue difficulty. It will therefore be found desirable in many cases either to simplify greatly the laboratory procedure by concentrating on only one or two of the possibilities, or to split the experiment up into two or three shorter experiments.

The measuring methods employed are all described in the main part of this book, and it is assumed the student performing the experiment will study the sections describing the apparatus and methods being employed. The theory involved in the various experiments is to be found either in this book or in the author's work "Radio Engineering." The experiments themselves are planned to supplement the theoretical treatment of radio and vacuum-tube phenomena given in "Radio Engineering."

The experiments have been planned in such a way as to require a minimum of special or expensive apparatus. In order to simplify the technique, most experiments involving tuned circuits are performed at audio frequencies. This eliminates many troubles that would otherwise be encountered, without altering the character of the phenomena involved.

In writing up the results it is assumed that the student will review those parts of this book and "Radio Engineering" that are related to the experiment, and that the usual explanations correlating experimental data with theory will be given. The exact details of the report will depend upon the particular apparatus used and will be affected by any modifications that may be made in the procedure.

### 1. BRIDGE MEASUREMENTS

**Object.**—To acquire familiarity with alternating-current bridges and the uses to which they may be put.

**Instructions.**—Make one or more measurements of the following types on a general-purpose alternating-current bridge. Use a Wagner ground and note the assistance it renders in increasing the sharpness of balance and removing body effects when the impedance of the unknown or standard arms is large.

*a.* Check resistance settings of a decade box.

*b.* Measure the capacity of a large condenser using a fixed standard.

*c.* Measure the capacity of a small condenser using the substitution method with a calibrated variable condenser as the standard.

*d.* Measure the inductance of an iron-cored coil using a fixed standard.

*e.* Attempt to measure the inductance of a broadcast tuning coil using a fixed standard, and then determine the inductance of the same coil using

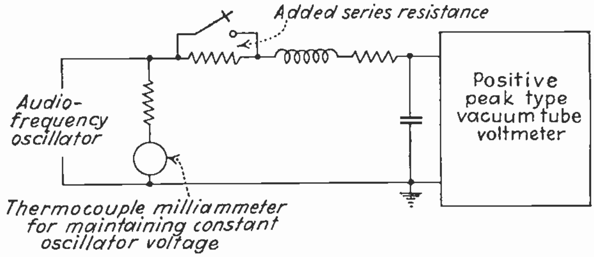



Fig. 199.—Circuit for Experiment 2.

a variable standard. Note the difference in the ease with which the balance is obtained in the two cases.

*f.* Measure the mutual inductance and calculate coefficient of coupling between two coils.

Determine the incremental inductance of an iron-core choke coil as a function of d-c current while keeping the alternating-current voltage across the coil constant with a vacuum-tube voltmeter arranged so that the d-c current is blocked off from the vacuum tube. This measurement should be made on a Hay bridge if one is available but can if necessary be made on a general-purpose bridge.

Make measurements on whatever other bridges are available.

## 2. RESONANT CIRCUITS—SERIES RESONANCE

**Object.**—To study the voltage and current relations existing in a series-resonant circuit.

**Instructions.**—Use a circuit consisting of an air-core inductance and a paper or mica dielectric condenser, and having a resonant frequency below 1000 cycles. Measure the circuit inductance, capacity, and resistance on a general-purpose bridge. Then set up the circuit as shown in Fig. 199 and, with a constant applied voltage, measure voltage across the condenser as a function of frequency for frequencies in the vicinity of resonance.

Repeat the run with an added known resistance approximately equal to the circuit resistance.

Plot the experimental results, together with theoretical curves of both magnitude and phase of voltage across condenser based upon the measured circuit resistance and $Q$. In calculating these theoretical curves assume that the experimental resonant frequency is the actual resonant frequency, calculate the rise of voltage at resonance from the circuit $Q$, and then use the usual working rules and the universal resonance curve to simplify the computations.

**Alternative Methods of Procedure.**—When suitable equipment is available for measuring small alternating currents over a considerable range of amplitudes, it is possible to measure circuit current as a function of frequency for constant applied voltage.

One can also modify the procedure by measuring impedance of the circuit as a function of frequency, using an alternating-current bridge, and then calculating the magnitude and phase of the current that would flow at different frequencies for constant applied voltage.

**Comments.**—The experiment is made by using a circuit that is resonant at audio frequencies because of the greater accuracy obtainable with simple experimental technique. The results obtained are similar in all respects to those obtained from circuits resonant at radio frequencies.

In making the calculations it will be noted that when the cycles off resonance exceed the resonant frequency divided by $Q/3$ to $Q/5$ the resistance has very little effect on the voltage across the condenser and so can be neglected.

## 3. RESONANT CIRCUITS—PARALLEL RESONANCE

**Object.**—To study the voltage, current, phase, and impedance properties of parallel resonant circuits.

**Instructions.**—Using the same resonant circuit as in Experiment 2, determine with a general-purpose bridge the magnitude and phase angle of the circuit impedance as a function of frequency (about resonance) when the circuit is connected for parallel resonance. This should be done for two or more values of circuit resistance.

Compute theoretical curves of circuit impedance from the constants of the circuit as measured in Experiment 2, and plot these upon the curve sheets with the experimental results. In making these computations, assume the experimental resonant frequency is the actual resonant frequency, and then obtain the impedance at resonance by the usual formula: Resonant impedance $= (\omega L)^2/R_s$. The impedance and phase angle at other frequencies can be determined with the aid of the customary working rules expressing sharpness of resonance in terms of $Q$, and by making use of the universal resonance curve.

**Alternative Procedure.**—Where equipment capable of measuring very small currents is available, one may carry out the experiment by applying a constant voltage to the parallel resonant circuit, and measuring the current flowing into the circuit as the frequency is varied about resonance. This gives an experimental curve of circuit impedance as a function of frequency. *In carrying out the experiment in this way it is absolutely essential that all* *harmonics of the supply oscillator be removed by placing a low-pass filter between the oscillator and the meter which is used to keep the input voltage constant.*

**Comments.**—In making calculations it will be noted that when the cycles off resonance exceed the resonant frequency divided by $Q/3$ to $Q/5$ the resistance has very little effect on the impedance and so can be neglected in ordinary calculations when this far off resonance.

Note the similarity of the impedance curves obtained in this experiment with the curves in Experiment 2.

## 4. COUPLED CIRCUITS—COUPLED IMPEDANCE

**Object.**—The study of the impedance which a tuned secondary couples into a primary circuit.

**Instructions.**—Inductively couple the same tuned circuit used in Experiments 2 and 3 to a primary circuit, and arrange a switch in series with the secondary circuit, as shown in Fig. 200. After determining the mutual inductance between primary and secondary coils, measure the coupled impedance. The coupled impedance is determined at any frequency by taking the difference between the impedance across the terminals of the primary coil when the switch in the secondary circuit is closed, and the impedance when the switch is open.

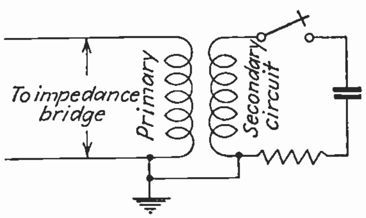



FIG. 200.—Circuit for measuring coupled impedance.

Plot experimental curves giving resistance and reactance components of the coupled impedance, and also the magnitude and phase of the coupled impedance, as a function of frequency. Calculate theoretical curves for these same quantities, using the known circuit constants, and plot upon the experimental curves.

**Comments.**—Note the similarity between the curve of coupled impedance in this experiment and the curves obtained with the same tuned circuit in Experiments 2 and 3.

## 5. COUPLED CIRCUITS—VOLTAGE AND CURRENT RELATIONS EXISTING WHEN PRIMARY AND SECONDARY ARE BOTH RESONANT CIRCUITS TUNED TO THE SAME FREQUENCY

**Object.**—Investigation of the behavior of coupled circuits when primary and secondary are both resonant at the same frequency, with special reference to band-pass effects.

**Instructions.**—Use the same secondary circuit as employed in Experiment 4, and use another identical circuit for the primary. Measure the inductance, capacity, and resistance of the individual circuits if this has not already been done. Apply a constant known voltage in series with the primary as in Experiment 2, and measure the voltage developed across this condenser in the secondary, also as in Experiment 2 (see Fig. 201). The should be done for frequencies about resonance, with the following coupling conditions:

*a.* Mutual inductance equal to value for maximum secondary current.

*b.* Mutual inductance 1.5 times value for maximum secondary current.

*c.* Mutual inductance equal to 3.0 times value for maximum secondary current.

*d.* Mutual inductance equal to 0.5 times value for maximum secondary current.

Plot all experimental results on one curve sheet. Calculate a theoretical curve for one case and plot it on the curve sheet with the experimental results.

Where band-pass effects are to be studied in detail, a desirable alternative (or addition) to the procedure given above is to determine the voltage across the secondary condenser as a function of frequency for constant

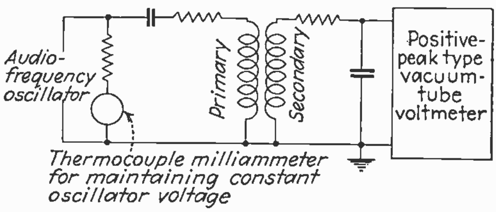



FIG. 201.—Circuit for Experiment 5.

coupling, but several values of circuit resistance (by adding equal known resistances to primary and secondary). The mutual inductance employed in this case should be sufficient to give pronounced double humps when no resistance is added to the circuits.

## 6. RADIO-FREQUENCY RESISTANCE

**Object.**—The measurement of resistance of radio-frequency circuits.

**Instructions.**—Measure the resistance of one or more tuned circuits suitable for use in broadcast receivers, making measurements at a number of frequencies in the broadcast band.

The procedure to use depends upon the equipment available. The resistance-neutralization method discussed in Sec. 21 is preferred; and if a bridge arrangement such as illustrated in Fig. 54 is not available, the arrangement of Fig. 52 can be set up, breadboard style. The only additional equipment required is then a broadcast receiver to detect the presence of oscillations.

An alternative procedure is the resistance-variation method also discussed in Sec. 21. This method requires a small oscillator capable of inducing a current of at least 100 ma in the circuit under test when loose coupling is employed. The induced current is customarily observed with a 120-ma thermocouple meter, the resistance of which must be subtracted from the measured circuit resistance. The resistance which must be added to the tuned circuit in this method of measurement can be conveniently supplied by a decade resistance box having good phase-angle characteristics. Several values of added resistance should be employed for each frequency, and the calculated circuit resistance for the different values averaged. Care must be taken to retune the circuit under test to resonance each time the circuit resistance is changed. The principal difficulty with the resistance-variation method is that the distributed coil capacity introduces errors, particularly when the tuning capacity is small. If the tuning condenser is calibrated and its setting read for each frequency of measurement, this distributed capacity can be calculated and allowed for as explained in Sec. 19.

Comments.—If the resistance-neutralization method is employed, it is possible to show the effect of shielding cans, neighboring metal or dielectric objects, etc., upon the circuit resistance and inductance. The change in inductance can be determined by using a calibrated tuning condenser and noting the change in capacity required to keep the resonant frequency of the circuit constant. The inductance used in calculating the resistance must, of course, be the actual inductance present, including the effect of shields, etc.

## 7. CIRCUIT CONSTANTS AT RADIO FREQUENCIES BY THE SUBSTITUTION METHOD

**Object.**—The measurement of circuit constants at radio frequencies, using the substitution method.

**Instructions.**—This experiment is carried out by using the same apparatus and tuned circuit that were employed in Experiment 6. (It makes no difference whether the resistance-neutralization or resistance-variation procedures are used.) Make measurements as follows:

*a.* Calibrate a variable condenser in terms of a calibrated variable condenser.

*b.* With the aid of a variable condenser having losses substantially independent of the capacity setting, determine the equivalent series resistance of one or more variable condensers as a function of their capacity setting, using the method described in Sec. 18. If time permits, also measure the equivalent series resistance of several small mica and paper dielectric condensers, and compare with an air condenser of the same capacity.

*c.* Measure the distributed capacity and the true inductance of a coil using the method illustrated in Fig. 48.

*d.* From the results of *c* obtain also the apparent inductance of the coil at several frequencies.

*e.* Determine the equivalent reactance and resistance of a radio-frequency choke coil at one or more frequencies, as discussed in Sec. 18.

## 8. CHARACTERISTIC CURVES OF TUBES

**Object.**—To obtain characteristic curves of the more important types of vacuum tubes, and to obtain the constants of the tube from the characteristic curves.

**Instructions.**—Obtain characteristic curves as follows:

*Diodes.*—Using a heater-type diode, determine plate current as a function of plate voltage for normal heater voltage, and for at least one subnormal heater voltage which shows voltage saturation at a relatively low plate voltage. Plot these curves in rectangular coordinates, and also replot the curve for normal voltage on logarithmic paper and determine the exponent $n$ in the equation $I_p = KE_p{}^n$.

*Triodes.*—Obtain curves of plate current as a function of grid voltage for several different values of plate voltage.

Obtain values of plate current as a function of plate voltage for several different values of grid voltage.

From the resulting curves determine values of $\mu$, $R_p$, and $G_m$ at one or more operating points.

*Screen-grid Tubes.*—Obtain curves of plate and screen currents, and also total space current, as a function of plate voltage for a fixed screen-grid potential and several values of grid bias.

Also determine plate, screen, and total space currents as a function of grid bias for a screen voltage recommended for use when the tube is operated as an amplifier, when the plate potential is (*a*) less than the screen voltage, and (*b*) considerably greater than the screen voltage.

Determine, in so far as it is possible to do so from the curves, the mutual conductance, amplification factor, and plate resistance of the screen-grid tube at an operating point suitable for amplifier operation.

*Pentodes.*—Repeat instructions given for screen-grid tubes, using only a radio-frequency pentode.

*Variable-mu Tubes.*—Using a variable-mu tube of the same general type as the pentode above, obtain a curve of plate current as a function of grid voltage for fixed screen-grid and plate potentials.

Plot upon the resulting curve a corresponding curve for a sharp cut-off pentode operating at the same screen-grid and plate potentials.

*Class B Tubes.*—Obtain curves of plate and grid current as functions of control-grid voltage for several values of plate voltage less than the normal supply voltage. Carry these curves in each case from cut-off to the largest positive grid potential that will not cause the allowable plate and grid dissipations to be exceeded.

Comments.—On each curve sheet mark the points which the manufacturer of the tube recommends for normal operation.

In obtaining data for curves, care must be taken to see that allowable electrode dissipations are not exceeded. This means that for high plate voltages the curves cannot be carried to the same current values as is permissible at low electrode potentials.

In carrying out the experiment it is instructive to use the same tube for all except the variable-mu runs. This can be done by use of a radio-frequency pentode, such as the 57 tube, with the electrodes connected to give different types of operation. Diode action is obtained by connecting all grids and the anode together to function as the plate electrode. Triode action is obtained when the screen and suppressor grids are tied to the plate, while screen-grid operation results when the suppressor grid is tied to the screen grid. Class B operation is obtained either by connecting all grids together for the control grid or by connecting together the two grids nearest the cathode and employing this combination as the control grid while connecting the suppressor to the plate. The advantage of using the same tube connected in different ways is that this shows the relations that the different types of operation have to each other.

## 9. TUBE COEFFICIENTS

**Object.**—To determine the amplification factor, mutual conductance, and plate resistance of tubes by bridge or null measurements.

**Instructions.**—Measure the amplification factor, mutual conductance, and plate resistance as a function of control-grid voltage, for constant plate voltage, using a triode tube.

With a pentode operated at plate and screen potentials suitable for amplifier use, measure mutual conductance as a function of control-grid voltage.

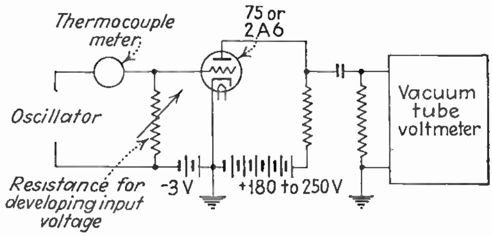



FIG. 202.—Circuit for measuring amplification obtained with resistance coupling.

In making the foregoing measurements the voltage-ratio method is preferable and should be used if the necessary equipment is available. When this method is employed, it is also possible to measure the amplification factor and plate resistance of the pentode tube. When the voltage-ratio bridge is not available, the measurements for this experiment can be made by setting up bridge circuits such as illustrated in Fig. 95.

## 10. RESISTANCE-COUPLED AMPLIFIERS

**Object.**—To design a resistance-coupled amplifier and measure its characteristics.

**Instructions.**—Before coming to the laboratory design a resistance-coupled amplifier utilizing either a 75 or a 2A6 high-mu triode. This design can be based upon data given in the tube manuals and should be such that the amplification falls off to about 70 per cent of its maximum value at 60 cycles.

Set up this amplifier in the laboratory, with a vacuum-tube voltmeter connected across the output as shown in Fig. 202. Measure the amplification as a function of frequency from 30 to 15,000 cycles by applying a voltage of variable frequency to the input, and adjusting this input voltage at each frequency until the output voltage as indicated by the vacuum-tube voltmeter has some predetermined value.

Repeat this measurement with half the design value of coupling condenser.

Repeat the measurement with the proper value of coupling capacity, but a capacity of approximately 100 $\mu\mu$f in shunt with the leak resistance.

Calculate a curve of amplification as a function of frequency for each case, and plot alongside of the experimental results. In making these calculations the load and grid-leak resistances can be measured on a direct-current bridge, the coupling condenser can be measured on a general-purpose bridge, and the stray shunting capacity determined from the fact that when the amplification at high frequencies drops to 70 per cent of the maximum amplification the reactance of the stray shunting capacity equals the resistance formed by placing grid-leak, load, and plate resistances in parallel. The tube constants at the operating plate current and grid bias are preferably obtained by a bridge measurement, but if this is not feasible the constants may be evaluated from the characteristic curves of the tube, or by using the incremental method. Fair results can even be obtained by using constants published in tube manuals.

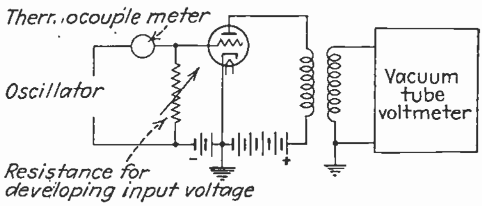



FIG. 203.—Circuit for measuring amplification obtained with transformer coupling.

**Alternative Procedure.**—An alternative (or additional) procedure is to use a radio-frequency pentode tube as the amplifier tube.

## 11. TRANSFORMER-COUPLED AMPLIFIER

**Object.**—To measure and calculate the amplification of a transformer-coupled amplifier.

**Instructions.**—Set up a transformer-coupled amplifier, applying the output voltage to a vacuum-tube voltmeter as shown in Fig. 203. Measure the amplification as a function of frequency from 30 to 15,000 cycles, using the same technique as employed in Experiment 10.

Repeat the measurements using an abnormally high negative bias on the grid of the amplifier tube so that the plate resistance of the amplifier tube is high.

Measure the tube constants as in Experiment 10, and obtain the transformer constants as described in Sec. 49. In determining the step-up ratio and effective shunting capacity the recommended procedure is to obtain a curve of voltage ratio as a function of frequency, by applying an input voltage directly across the transformer primary while leaving the wiring and vacuum-tube voltmeter on the secondary side unchanged.

Calculate amplification curves for the two runs and plot alongside of the experimental results.

## 12. FREQUENCY CHARACTERISTICS OF CLASS A POWER AMPLIFIERS

**Object.**—To measure and calculate the way in which the amplification of a Class A power amplifier varies with frequency.

**Instructions.**—Set up a Class A power amplifier using the circuit arrangement shown in Fig. 204. The output transformer should be one designed to be used with the tube and circuit employed, and the load should be that for which the transformer is designed.

Measure the variation of output voltage (or current) with frequency when some convenient input voltage having a crest value of not over one-half of

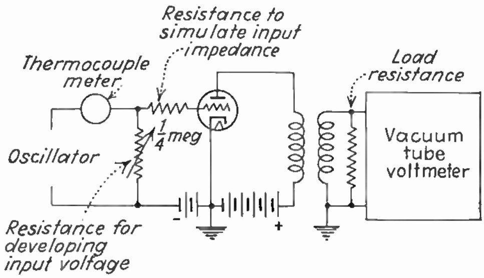



FIG. 204.—Circuit for measuring frequency distortion in a Class A power amplifier.

the grid bias is applied. The output can be determined either by a vacuum-tube voltmeter shunted across the load or by a thermocouple milliammeter placed in series with the load. In the latter arrangement the resistance of the thermocouple must be considered as part of the load and allowed for. Repeat for load resistances one-half and twice the recommended values.

Measure the tube and transformer constants and calculate a theoretical curve of amplification as a function of frequency. The tube constants required are amplification factor and plate resistance at the operating point, while the important transformer constants, as indicated in Sec. 49, are direct-current resistance of primary and secondary, leakage reactance, incremental inductance of the primary winding with the appropriate d-c passing through the primary in the correct direction, and the transformer ratio.

## 13. AMPLITUDE DISTORTION AND LOAD LIMIT IN CLASS A POWER AMPLIFIERS

**Object.**—To measure the amplitude distortion in a Class A power amplifier, and to determine experimentally the factors that affect the load limit.

**Instructions.**—Use the experimental setup of Experiment 12, supplemented by means for measuring the distortion contained in the output. It is also desirable to provide a filter between the oscillator and the amplifier input in order to insure a sinusoidal exciting voltage. The distortion can be measured by whichever method described in Chap. VI is the most convenient.

With the same load conditions as employed in Experiment 12, and a frequency of 400 to 1000 cycles, determine fundamental, second-, and third-harmonic currents, and the grid current, as a function of input voltage, carrying the input voltage to well beyond the normal full-load value.

Arrange an oscillograph so that the shape of the output wave can be observed. This may be either an electromagnetic oscillograph with an element in series with the load or a cathode-ray oscillograph deflected electrostatically by the alternating voltage between plate and cathode and having a linear sweep circuit. Observe the character of the distortion as the amplifier becomes overloaded: first, when the tube has normal plate and grid-bias voltages, and the load resistance is 50, 100, and 200 per cent of the recommended values, and, second, with the load resistance at the recommended value, but with grid-bias voltages such that the d-c plate current is 50, 100, and 150 per cent of the normal values.

Obtain a set of characteristic curves for the power tube, and work out the dynamic characteristic for the normal operating point, using the recommended load resistance and a signal voltage having a crest value equal to the grid bias.

## 14. MULTISTAGE AMPLIFIERS

**Object.**—The determination of the characteristics of a multistage audio-frequency amplifier, including the over-all amplification, the amplification of each individual stage, and the effects produced by a common plate impedance, and by improper by-passing of bias resistors.

**Instructions.**—With a multistage amplifier operating under normal conditions determine the over-all amplification and the amplification of each individual stage, as a function of frequency, using the method described in Sec. 46.

Repeat the measurements when the by-pass condenser around one of the self-bias resistors is reduced to a value such that the amplification at low frequencies is appreciably affected.

Repeat the measurements after reducing the plate-circuit filtering to the point where there is appreciable regeneration present (*i.e.*, until just before oscillations start).

Reduce the filtering in the plate circuits until oscillations take place, and note the character of these oscillations.

**Comments.**—This experiment requires the use of a suitable amplifier already built up in a way that will allow circuit constants to be readily altered. It should also be possible to measure plate currents, grid-bias voltages, etc., and to test the transformers. The amplifier can consist of one stage of triode transformer-coupled amplification, one stage of resistance coupling, using either a pentode tube or a high-mu triode, and a power stage. The last may be push-pull if desired.

This experiment can also be used to demonstrate some of the characteristics and advantages of push-pull power amplification. To do this the power stage should be arranged so that it can be operated either as a push-pull or single-tube amplifier. Comparison can then be made of the two methods of operation showing particularly the advantages of push-pull amplification in reducing regeneration from a common plate-circuit impedance, and eliminating second-harmonic distortion.

### 15. CLASS B AUDIO-FREQUENCY POWER AMPLIFIERS

**Object.**—Study of the characteristics of a Class B audio-frequency amplifier, with special emphasis upon the factors controlling distortion.

**Instructions.**—Arrange a Class B audio-frequency power amplifier as shown in Fig. 205, with a positive-peak vacuum-tube voltmeter to read the maximum positive potential reached by the grid, and a trough vacuum-tube voltmeter to indicate the minimum positive potential reached by the plate. The input transformer should be of the usual type employed in driving the amplifier, and the resistance $R_p$ should be of the proper value

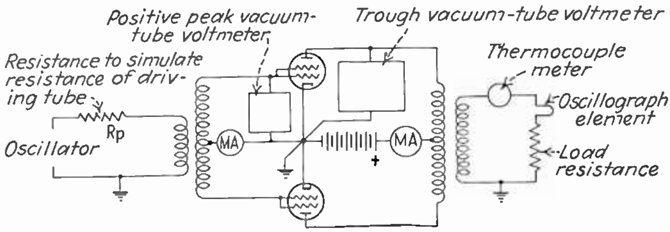



FIG. 205.—Circuit for studying operation of Class B audio-frequency power amplifier.

to stimulate the plate resistance of the driving tube. The wave shape of the output is observed with the aid of an electromagnetic oscillograph element in series with the load resistance, or with the aid of a cathode-ray oscillograph having a linear sweep circuit.

With normal values of $R_p$ and load resistance, and an input signal having a frequency of 400 to 1000 cycles, increase the input voltage until the rated output power is obtained, at which point measure the maximum positive grid potential and the minimum positive plate potential. Then without changing the input voltage, increase the load resistance until the wave is visibly distorted, and read maximum positive grid potential and minimum positive plate potential. Likewise note any change in output power, d-c plate current, and d-c grid current.

Restore the normal load resistance and, with the input voltage unchanged, double $R_p$. Read maximum positive grid potential, minimum positive plate potential, power output, and note effect on d-c grid and plate currents.

For each case sketch oscillograms approximately to scale showing the way in which the instantaneous plate and grid currents and voltages vary during the cycle.

### 16. TUNED AMPLIFIERS

**Object.**—Study of the factors involved in tuned amplifiers.

**Instructions.**—Arrange a tuned amplifier as shown in Fig. 206 using the same tuned circuit as that employed in Experiment 2. With the input voltage kept constant, measure the way in which the output voltage varies with frequency about resonance when the coupling between primary and secondary is such as to give (*a*) maximum possible amplification, (*b*) maximum possible mutual inductance, (*c*) about one-half of the coupling for maximum amplification.

Measure mutual inductance accurately for each case, and also determine the amplification factor and plate resistance of the tube at the operating point employed.

Calculate theoretical amplification curves and plot alongside of the experimental results. In making these computations assume that the experimentally determined resonance frequency is the actual resonant

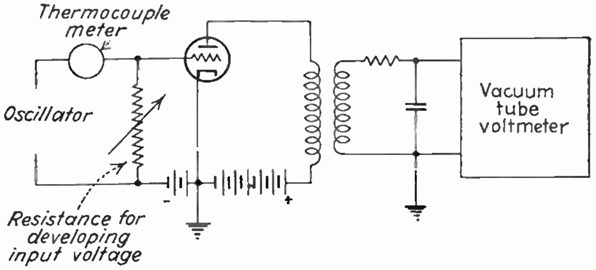



Fig. 206.—Measurement of the characteristics of tuned amplifiers.

frequency, and then for each case calculate the amplification at resonance and the effective $Q$ of the amplification curve. Determine the remainder of the curve with the aid of the universal resonance curve.

## 17. INPUT ADMITTANCE IN AMPLIFIERS AND ITS NEUTRALIZATION

**Object.**—To investigate the nature of the input admittance of a tuned amplifier.

**Instructions.**—Use the tuned amplifier employed in Experiment 16 with the coupling giving maximum possible amplification. Artificially increase the grid-plate capacity by connecting a fixed condenser of about 50 $\mu\mu$f between grid and plate. Arrange a general-purpose bridge so that the input impedance of the tube can be measured by the substitution method. The experimental setup is shown in Fig. 207.

Measure the input admittance of the amplifier as a function of frequency about resonance, determining both magnitude and phase angle.

Next neutralize the input admittance by connecting the condenser $C_n$ shown dotted in Fig. 207, and adjusting as necessary. In neutralizing it may be necessary to reverse the primary terminals in order to obtain the correct phase relation. When the neutralization has been completed, remeasure the input admittance of the amplifier as a function of frequency.

If it is desired, one may overneutralize the amplifier and repeat the measurements again.

Plot experimental curves of input capacity and resistance for each case.

go far enough into the positive grid region to permit deriving theoretical curves of grid- and plate-current waves, knowing the grid- and plate-voltage waves. One can then sketch out a complete set of oscillograms such as shown in Fig. 124 and can determine the theoretical d-c plate and grid currents, power input and power output, and efficiency, by counting squares under each curve or using a planimeter and averaging over the cycle.

**Comments.**—The success of this experiment depends largely upon obtaining the proper proportions for the tuned circuit, but when all the details are properly worked out this is one of the most effective experiments that can be performed in a university laboratory.

The tuning inductance must be air-cored and, in order to simulate circuit impedances actually used at radio frequencies, should have a reactance of about 600 ohms when direct coupling such as shown in Fig. 208 is employed. The Q of the tuned circuit with no added resistance should exceed 25, and the added resistance used to simulate load should never be great enough to make the circuit Q fall below 10. The tuning condenser is required to stand a crest-alternating voltage that is only slightly less than the d-c plate potential, and so must have a high voltage rating. Tuning can be accomplished by varying the inductance of the tuned circuit, or by varying the tuning capacity in steps of about 2.5 per cent. Arrangements that have been found satisfactory include the use of a 50-watt tube operated at 500 volts plate potential (which is half of the normal voltage rating), an air-cored inductance of 0.4 henry which is tuned to 250 cycles by use of a fixed condenser supplemented by a condenser variable in small steps. A self-contained two-element portable oscillograph is found to work very satisfactorily.

## 19. MODULATED AMPLIFIERS

**Object.**—Study of plate-modulated and grid-modulated Class C amplifiers, with special emphasis upon the factors controlling the linearity of modulation.

**Instructions.** *Plate-modulated Class C Amplifiers.*—Use the same arrangement of apparatus as in Experiment 18. Determine the grid bias which is required to bring the plate current to nearly zero with no exciting voltage, and then use twice this value as the operating grid bias. Adjust the exciting voltage and plate load resistance so that the maximum positive grid potential is slightly less than the minimum positive plate potential, while obtaining at the same time a reasonable power output. Finally, simulate modulation by varying the direct-current plate potential from zero to approximately twice normal value and obtain amplitude of output voltage or current, and also the d-c plate current, as a function of direct-current plate voltage. The grid exciting voltage must be constant during this procedure and may have to be readjusted slightly as the plate voltage changes. The amplitude of the output can be most accurately determined by measuring the minimum positive plate potential for each point and subtracting from the direct-current plate potential to obtain the alternating plate-cathode voltage (which in turn is proportional to the output current).

Reduce the load resistance to zero, readjust the excitation until the maximum positive grid potential is slightly less than the minimum plate potential; and repeat the run. Note the increased linearity in the modulation that is obtained at the expense of a reduction in power output.

*Grid-modulated Amplifiers.*—Use the same arrangement of apparatus as for plate modulation, except that the positive peak grid-voltage meter may be dispensed with. Determine the grid bias required to reduce the plate current very nearly to zero when there is no exciting voltage, and then adjust the exciting voltage until traces of grid current begin to appear. With this excitation kept constant, obtain the relation between grid bias and output voltage or current, first, for lowest possible resistance in series with the tuned output circuit, and, second, with considerable added load resistance in series with the tuned output circuit. Carry these curves to the point where all output disappears, and note the difference in linearity in the two cases.

**Comments.**—It will be observed that in this experiment the audio-frequency exciting voltage represents the carrier, while an adjustable direct-current voltage simulates the modulating voltage. This procedure greatly simplifies the experimental procedure and yet gives results exactly the same as would be obtained by using a radio-frequency carrier voltage.

The tubes employed must be operated sufficiently far below their rated capacity so that at the crest of the modulation cycle (when the current or voltage output is twice the normal value) the tube has only just reached its full rating. Otherwise the tube may be damaged.

Any oscillator or Class C amplifier tube may be used as a Class C modulated amplifier, but for grid modulation the tube must have a low amplification factor if appreciable output is to be obtained without driving the grid positive.

## 20. LINEAR AMPLIFIERS

**Object.**—To determine the performance of a linear amplifier.

**Instructions.**—Use the same setup as in Experiment 18, except arrange the positive peak meter in the grid circuit so that it can be switched to read either maximum positive grid-filament voltage or peak value of exciting voltage. Adjust the grid bias to "projected cut-off" (*i.e.*, to a bias corresponding to that for which a projection of the straight-line part of the tube characteristic cuts the zero current axis), and determine d-c plate current and the output voltage (or current) as a function of exciting voltage for (*a*) no added resistance in series with the tuned load circuit, and (*b*) the same added resistance as used in the reference condition in Experiment 18. Carry these curves to the point where the grid begins to draw a disproportionate fraction of the total space current.

Repeat for the same added resistance in the tuned circuit as used above, but with grid-bias voltages twice and half the projected cut-off values.

Note the relation between the minimum plate voltage and maximum grid potential at the point where the linearity ceases, and also note relative linearity, power, and efficiency obtainable for the different conditions.

**Comments.**—It will be noted that varying the exciting voltage from zero to a large value is equivalent to applying a modulated wave to the amplifier, and the linearity between output and input represents the fidelity with which the modulation envelope is reproduced in the output.

## 21. POWER OSCILLATORS AND MODULATED OSCILLATORS

**Object.**—To adjust a power oscillator to good efficiency and reasonable output, and to study some of the properties of the oscillator.

**Instructions.**—Connect an oscillator as shown in Fig. 209, and tune to generate some preassigned radio frequency (using a wavemeter). Adjust grid, plate, and filament taps to develop the maximum output available with an efficiency of about 50 per cent or more. In making these adjustments keep in mind the fact that the voltage and current relations in a power oscillator are similar to those in a Class C power amplifier.

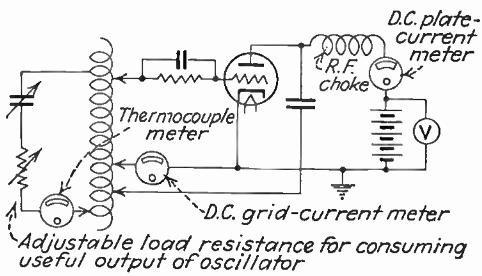



FIG. 209.—Circuit for studying vacuum-tube oscillator.

After the oscillator has been adjusted, note relative effects on frequency of adjusting plate, grid, filament, and tuning condenser taps, using a wavemeter as the frequency-measuring device. Note the effect on the oscillator plate and grid currents of closely coupling the wavemeter to the oscillator and tuning through resonance. Vary the resistance in series with the tuning condenser and observe the results.

Vary the direct-current plate voltage from zero to well beyond the voltage used in the first part of the experiment, and measure the relative amplitude of oscillations and the d-c plate current as a function of direct-current plate voltage. The relative amplitude of oscillations can be measured on a vacuum-tube voltmeter or a thermocouple milliammeter associated with the oscillating circuit. From the results determine the amount of modulating power required for 100 per cent modulation, and discuss the amount of distortion that could be expected.

## 22. DETECTORS AND VACUUM-TUBE VOLTMETERS

**Object.**—Experimental study of the fundamental characteristics of detectors and vacuum-tube voltmeters.

**Instructions.**—Determine the change in d-c plate current as a function of the amplitude of a sinusoidal applied voltage for (*a*) a grid-leak power detector of normal design, (*b*) plate detector biased approximately to cut-off, (*c*) plate detector with bias considerably less than cut-off, and (*d*) a plate detector with bias considerably greater than cut-off; (*e*) also measure d-c current flowing through the leak resistance of a typical diode detector as a function of applied voltage.

In making these measurements it is permissible to use an audio-frequency voltage rather than a radio-frequency voltage if the condenser shunting the leak in the grid-leak and diode detectors is increased in size until it is an effective by-pass for the frequency employed.

In interpreting the results it is to be kept in mind that a modulated wave applied to the detector is equivalent to a sine wave of varying amplitude, so that, if the modulation frequency is low enough to permit the rectification process to follow the variations of the modulation envelope, it is possible

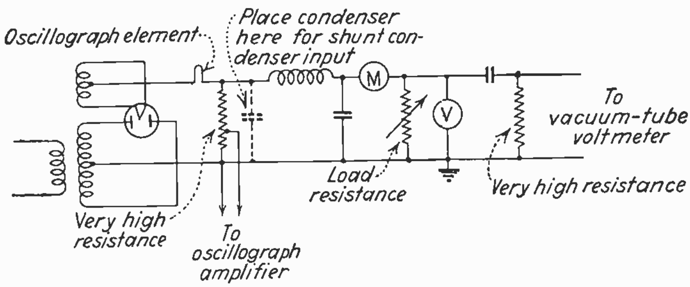



FIG. 210.—Circuit for rectifier-filter experiment.

to deduce the load limits, the distortion, and the output from the data obtained.

Discuss the results with respect to distortionless power detection of broadcast signals. Also discuss the results for plate detection from the point of view of vacuum-tube voltmeters.

**Alternative Procedure.**—When equipment is available for producing a sinusoidally modulated wave, the scope of the experiment can be greatly extended by applying carrier voltages of various amplitudes, and with degrees of modulations up to 100 per cent, and observing the wave shape of the detector output in an oscillograph, as well as measuring the output voltage with a vacuum-tube voltmeter.

## 23. RECTIFIER-FILTER SYSTEMS

**Object.**—A study of the voltage and current relations existing in condenser-input and inductance-input rectifier-filter systems.

**Instructions.**—Connect up an inductance-input filter as shown in Fig. 210 using a low-resistance tube such as the 5Z3 for the rectifier. Place one element of an electromagnetic oscillograph in series with the input inductance, and connect a second element in the output of an amplifier arranged to show the voltage wave across the filter input.

Vary the load resistance from open circuit to well beyond full load, and read direct-current voltage and current at the load, and ripple voltage across load, as a function of load resistance.

Note the changes occurring in the oscillograph wave forms as the load current becomes very small. Sketch wave forms for several typical conditions.

Next place a shunt condenser across the input to the first inductance, and again determine direct-current voltage and current in load, and ripple voltage across load, as a function of load resistance. Also sketch the oscillograph wave forms for typical cases.

Determine the effect upon the ripple voltage across load of (*a*) doubling the size of condenser shunting the load, and (*b*) adding an additional section of filter before the load.

**Comments.**—The input inductance used should have such a high current rating and large air gap that there is practically no change in the incremental inductance up to the maximum direct current used in the test. This simplifies the experiment by avoiding excessive saturation, and the resulting instability of the incremental inductance. When a non-saturating inductance is employed, it is possible to calculate with fair accuracy the hum output voltage to be expected for the inductance input case, following the methods described in Sec. 95 of "Radio Engineering." The performance with condenser input may also be calculated by formulas given in "Radio Engineering" but as these formulas neglect the leakage inductance of the transformer, and the leakage inductance is of considerable importance, the results cannot be expected to be more than qualitative.

## 24. CHARACTERISTICS OF RADIO RECEIVERS

**Object.**—Determination of the performance of a typical radio receiver.

**Instructions.**—Measure the sensitivity, fidelity, and selectivity of a broadcast receiver at carrier frequencies of 600, 1000, and 1400 kc, using a standard-signal generator, and following the standard test procedure. If the receiver has automatic volume control, either the selectivity curve must be taken by the two-signal generator method, or the automatic volume control must be disconnected so that the receiver operates at its greatest sensitivity irrespective of the signal amplitude.

**Supplementary Procedure.**—When it is desired, this experiment may be expanded to include measurements of load limit, hum, automatic volume-control characteristics, etc., as well as a determination of the performance of each individual part of the receiver as discussed in Sec. 52.

## 25. RADIO TRANSMITTERS

**Object.**—Study of the operation and adjustment of a radio transmitter.

**Instructions.**—This experiment requires the use of an amateur radio transmitter, and the procedure to be followed depends upon the apparatus available. If the transmitter is a crystal-controlled code outfit, the experiment can involve tuning up the entire transmitter, determining the power output delivered to a dummy antenna and the plate efficiency of the output stage, checking the signals with monitor, etc. With a telephone transmitter one can in addition determine the linearity of modulation by using an oscillograph or a modulation meter, and the linearity of the linear amplifiers if these are employed.

An alternative procedure is to set up and adjust a small crystal-controlled telephone transmitter having 5 to 15 watts of carrier power.

## 26. RADIO-FREQUENCY TRANSMISSION LINES

**Object.**—A study of the current distribution and impedance of resonant and non-resonant lines.

**Instructions.**—Using a radio-frequency transmission line having a length in excess of one wave length, measure the sending-end impedance of the line as a function of load resistance while keeping the frequency constant, using the substitution method of measurement.

Measure the sending-end impedance as a function of frequency over a two to one frequency range for load resistance of (*a*) characteristic impedance, (*b*) about 10 per cent of the characteristic impedance, (*c*) about ten times characteristic impedance, (*d*) capacitive load having a reactance equal to the characteristic impedance.

Determine the relative current distribution for each of the foregoing cases, using one of the methods shown in Fig. 146. In plotting these results, express position in terms of wave lengths from the load end of the line.

**Comments.**—The exact details of this experiment will depend upon the facilities available with respect to both apparatus and space. An outdoor transmission line a few hundred feet long and within arm's length of the ground is preferable. The transmitter can then be operated at a wave length of 10 to 20 meters. If this arrangement is not possible, an alternative is to run a transmission line around or across a large room, back and forth several times if necessary. The driving oscillator should then operate at a somewhat shorter wave length (higher frequency) than for the case of an outdoor line in order that the line length expressed in wave lengths will be about the same.

## 27. ANTENNA MEASUREMENTS

**Object.**—Investigation of antenna characteristics.

**Instructions.**—The procedure to be followed in experiments on antenna characteristics depends so much upon the measuring apparatus and antenna facilities available that it is impossible to give blanket instructions. All that can be done is to suggest types of measurements that can be made. These include: (*a*) impedance between an antenna and ground as a function of frequency over a wide enough frequency range to show curves such as those of Fig. 144, (*b*) current distribution over a corresponding frequency range, (*c*) distribution of field strength about an antenna, and (*d*) directional characteristics of a simple antenna array.

**Comments.**—Experiments involving directional characteristics or antenna arrays, such as (*c*) and (*d*) above, can be most satisfactorily carried out at wave lengths of a few meters. At these wave lengths it is even possible to set up a directional antenna system, such as a diamond (or rhombic) antenna, a V (or partially folded long wire), or an array of several half-wave antennas, that is physically small enough to be investigated inside a large laboratory room. The ground for such an antenna can be provided by a copper- or chicken-wire screen.

Directional characteristics in the vicinity of the antenna can be obtained by the use of a simple radio receiver consisting of a short rod antenna coupled to a tuned circuit that operates a vacuum-tube voltmeter. The rod antenna should have a coupling coil at its middle that is electrostatically shielded from the tuned circuit in order to preserve symmetry with respect to ground. By arranging the antenna rod so that it can be rotated on a universal joint it is possible to investigate polarization of the waves radiated from the antenna.

If the antenna in either (*c*) or (*d*) is an ungrounded type, it is possible to investigate the effect of height above ground by taking the directional characteristics in a vertical plane in the direction of the main beam, for two heights above ground.

It must be remembered that antennas supplied with power will radiate radio signals. Such experiments must therefore be carried out in accordance with the regulations governing the transmission of radio signals. In particular, the station must be licensed, and a licensed operator must be on duty.

## 28. FIELD STRENGTH OF RADIO WAVES

**Object.**—Measurement of the field strength of a radio wave.

**Instructions.**—Measure the field strength of a number of broadcast stations, using the substitution method as illustrated in Fig. 134.

## 29. WAVE PROPAGATION

**Object.**—The study of various aspects of wave propagation.

**Instructions.**—The procedure to follow in studying wave-propagation phenomena will depend upon the facilities available. Among the things that can be observed are variation of field strength with time (fading), quality distortion and the point upon the fading cycle at which quality distortion is normally worst, round-the-world echo signals, reflections from the Kennelly-Heaviside layer, polarization characteristics of incoming signals, etc.

Fading can be readily observed and evaluated with the aid of a receiver having automatic volume control, and using a microammeter to indicate the current in the automatic volume-control system. If everything else is kept constant, this current is a measure of the signal strength, so that fading can be observed by recording the changes in meter reading with time. A calibration giving the relation between meter reading and actual signal strength can be obtained with the aid of a signal generator.

Round-the-world echo signals are always audible at certain times of the year at certain hours of the day, by using an ordinary short-wave code receiver.

If an amateur transmitter capable of being modulated in very short pulses is available, it is a relatively simple matter to observe Kennelly-Heaviside layer reflections, by using a cathode-ray tube to observe the received signals, and synchronizing both the transmitted pulses and the cathode-ray sweep voltage with the same 60-cycle source.

## 30. THE RADIO COMPASS

**Object.**—To study direction finding, using a loop antenna.

**Instructions.**—Set up a loop antenna receiving system, using a sensitive radio receiver, and arrange the coupling between loop and receiver so that there are no unbalances. Determine the bearing of several near-by broadcast radio stations and compare with the true bearing. If possible to do so, use an auxiliary vertical antenna to determine the sense of the bearing in one or two cases.

Obtain bearings upon the most distant broadcast signal heard in the daytime, and then take bearings again in the evening upon the same stations. Note particularly the erratic and variable character of the night bearings.

If time is available, deliberately unbalance the loop antenna with respect to ground and observe the effects upon the measurements.

## 31. MEASUREMENT OF FREQUENCY

**Object.**—The measurement of frequency at audio and radio frequencies.

**Instructions.** *Audio-frequency Measurements.*—Check the calibration of an audio-frequency oscillator at several points by using either a Wien or a resonance bridge.

Using a cathode-ray tube, a fixed-frequency source (such as 60 cycles), and a variable-frequency oscillator, produce Lissajous figures and, if time is available, either gear-wheel or spot-wheel patterns, for different frequency ratios.

*Radio-frequency Measurements.*—The use of a wavemeter to measure frequency should be included here if the wavemeter has not already been used in previous experiments.

Calibrate an oscillating-detector type radio receiver by the use of harmonics of a fixed-frequency oscillator. These can be harmonics of a crystal oscillator, or they can be harmonics of a multivibrator which is synchronized with some standard frequency.

**Supplementary Procedure.**—Set up a multivibrator with circuit proportions such that the natural frequency of oscillation is about one hundred cycles, and if possible arrange so that the wave form can be observed. A magnetic oscillograph in conjunction with an oscillograph amplifier is suitable, or a cathode-ray tube with a linear sweep circuit will do by connecting one pair of plates between ground and the grid of one of the multivibrator tubes. Arrange a coupling tube so that the pitch of the multivibrator oscillations can be observed in a telephone receiver. Inject into the multivibrator circuits an oscillation having a frequency of about ten times the multivibrator frequency. Gradually increase the amplitude of the injected voltage and note the changes in wave form and multivibrator frequency that are produced.

**Comments.**—The most satisfactory procedure for demonstrating the use of harmonics in frequency calibration at radio frequencies will depend upon boxes provided with a bakelite cover plate, upon which can be mounted binding posts, and also one or two seven-prong tube sockets wired to make available a variety of voltages. Equipment such as oscillators, amplifiers, and radio receivers can then be provided with cords terminated in plugs that go into the tube sockets on the outlet boxes. This avoids the inconvenience and cost of individual batteries or the expense of providing each piece of equipment with its own rectifier-filter system. A battery system providing 250 volts tapped at 140 and 90 volts for plate supply, a heavy-current 6-volt battery, and grid-bias voltages of −24 and −48 volts is adequate for ordinary situations. The battery voltage can be kept at the rated value by continuous trickle charging. The initial cost of batteries is appreciable; but when it is realized that the current drain is so small that the lightest storage cells are satisfactory and will give a life of the order of 10 years, the yearly cost is far less than the savings made possible.

Vacuum-tube voltmeters are used so frequently in communication work that they should be built up in permanent form. Such units provided with self-contained batteries (45-volt B battery, C battery, and two No. 6 dry cells for heating the filament of a dry-cell tube) can be used almost as one would use an ordinary voltmeter and when once available will be found indispensable.

Some of the equipment needed by a communication laboratory must be built because suitable commercial apparatus is not available. In other cases it may be necessary to build one's own equipment or do without because of lack of funds to purchase commercial apparatus. This is particularly true of university laboratories, where student help with shop experience is often available even when there is very little cash to spend. As an indication of what is possible in the construction of equipment, an incomplete list of apparatus that has been built in the communication laboratory at Stanford University under the author's supervision is given below.

Beat-frequency oscillator.

Resistance-stabilized audio-frequency oscillators (four in all).

Incremental inductance bridge (as illustrated in Fig. 39).

Dynatron bridge (see Fig. 54).

Standard frequency outfit (similar to that shown in Fig. 76).

Audio-frequency microvolter.

Vacuum-tube voltmeters with self-contained batteries.

Peak and trough vacuum-tube voltmeters (Figs. 18*a* and 19*b*).

Tube-testing equipment, such as illustrated in Fig. 99, together with unit for measuring tube characteristics by the voltage-ratio method (see Fig. 98).

Standard signal generator of the mutual-inductance type (see Fig. 121).

Cathode-ray oscillograph unit.

Linear sweep circuit for cathode-ray tube.

Decade resistance, inductance, and capacity boxes.

Standard voltage generator for calibrating vacuum-tube voltmeters (see Fig. 17).

Amplifiers, including oscillograph amplifier, microphone and bridge amplifiers, and an amplifier capable of delivering over 25 watts of undistorted audio power for use in Class C amplifier and similar experiments.

Wavemeters.

In addition a number of direct-current and rectifier voltmeters and milliammeters have been mounted in boxes, provided with 8 to 11 ranges, and completely fused. Every multiple-range meter of this sort has been found to be equivalent to from two to four single-range meters.

When equipment is being built, a small shop is almost a necessity. The tools needed are drill press, grinder, vise, hacksaw, tin snips, drills, center punch, reamers, screw drivers, etc. A portable electric drill is also desirable. The sum of $100 carefully spent will provide everything essential.

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