EXPERIMENTAL RADIO ENGINEERING

BY

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PREFACE

To the best of the author's knowledge there is no text book at present available which is directed specifically to the teaching of radio principles in the laboratory. There are available some brief outlines of various tests on radio apparatus, but in such books the keen student will find no satisfaction. One learns but little of the fundamental principles of radio in adjusting a radio receiver.

A comprehensive text for the radio laboratory should be laid out with the idea of emphasizing the principles involved in the performance of radio apparatus, rather than the specific operation of such. A factory test man can generally carry out the adjustment of a radio receiver in routine fashion just as well as or better than the graduate of the best engineering school; it is of little value to the engineering student to spend much time in work of this nature. Rather should his efforts be directed to mastering the relationships of the multitudinous factors which affect the performance of such a receiver. Some one has the task, therefore, of dissecting into its simple constituents, receiving and transmitting apparatus, and laying out such tests as shall bring out the characteristics of these component parts. Then, in so far as the performance of one part affects the operation of another, other tests must be laid out to show these mutual relationships.

Such is the viewpoint and purpose of the author in writing this laboratory manual. It is anticipated that the laboratory work will be carried out in parallel with classroom work which uses some such text as the author's "Principles of Radio Communication."

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INTRODUCTION

Reasons for Performing Some of the Tests at Low Frequencies.—When endeavoring to verify experimentally the laws underlying the performance of radio apparatus it is advisable to eliminate, in so far as possible, errors of variable and unascertainable character. The errors generally increase in importance as the frequency is raised, and at frequencies of a million or more, are frequently of such a character that even the most skilled radio experimenter finds difficulty in checking theoretical and experimental results.

As an illustration of this let us consider the apparently simple task of measuring the amount of amplification of an ordinary broadcast receiver. The identical receiver, left in the same place in the laboratory, and tested with the same instruments, shows variations from day to day not of a few per cent but of hundreds of per cent! In Fig. 1 this fact is illustrated; the amplification of this special receiver when tested every day for a month gave results which increased and decreased irregularly from day to day, as much as covered by the shaded area. This shows that when adjusted to receive different frequencies the voltage amplification varied by different amounts, for the higher frequencies changing as much as

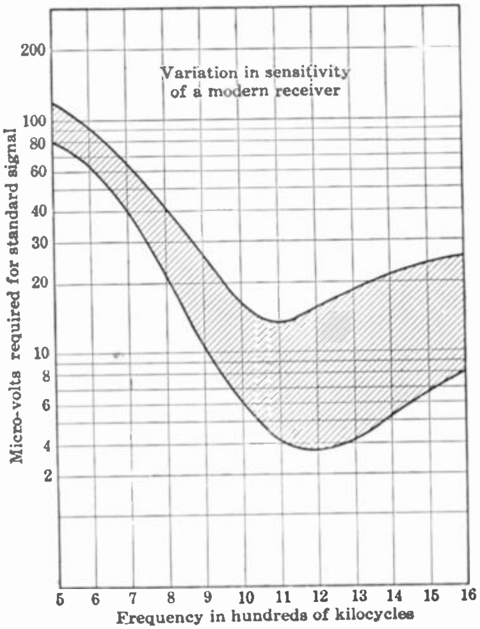


FIG. 1.

four times. Later tests seem to show that this difference in performance was due to moisture condensed on parts of the apparatus. A correlation was found between amplification, humidity, and temperature, with a time lag between cause and effect of several days!

Evidently the beginning experimenter cannot be expected to avoid, or account for, such effects. At the low frequencies used in power machinery such errors are not likely to occur, and at these frequencies, it is easy to obtain a close agreement between theory and experiment, for many of the circuits used in radio apparatus.

Meters for Use in the Radio Laboratory.—As will be evident from examination of the list of experiments outlined in this text, some of them are to be carried out at power frequencies, some at various frequencies in the audible range, and many of them at radio frequencies.

For the power frequency experiments the ordinary type of alternating current ammeter or voltmeter is available, notably the iron vane type such as the Weston Model 155. Besides being cheap and rugged they are reasonably accurate for frequencies up to a few hundred cycles per second.

In using meters of this class the student is likely to acquire an attitude of faith in the ability of his meters to stand overloads and rough usage which will be rudely shaken when he begins to use the more delicate and sensitive types of meters employed in radio measurements. For example, if 10 amperes is put through a 5-ampere iron vane meter, or 200 volts impressed on a voltmeter of 100-volt range of this type, no appreciable harm is done the meter, if the overload is of short duration. The finger of the meter will be completely off the scale and the student is at once aware of the overload.

A 100 per cent overload on a hot wire meter, or thermocouple meter, however, is always disastrous, and the meter is ruined. Furthermore, many times the response of the meter is so sluggish that the meter is burned out before the finger has had time to move even to full scale reading. A change in the setting of the variable condenser in a radio circuit, of only one of two divisions, may bring about such a large change in current as to burn out a meter which, before the setting of the condenser was changed, was reading only a small percentage of full scale. In some circuits the mere approach of the observer, to take a meter reading, may change the resonance condition in the circuit sufficiently to burn out a thermocouple which, before the observer drew near, was reading well down on its scale.

The student beginning to work in the radio laboratory cannot observe too much caution in the use of meters of the hot wire and thermocouple types.

Ordinary Meters Not Suitable for Radio Frequencies.—The question might well be asked—why not use the rugged meters of the iron vane type, or dynamometer type, in making radio measurements? The answer is that these meters, if they indicate at all at radio frequencies, will be entirely incorrect in their indications, for one or more of several reasons.

The iron vane type of meter utilizes the repulsive force between iron vanes, magnetized by a coil through which is flowing the current to be measured. For a given strength of current the amount of magnetic flux set up in the iron vanes decreases as the frequency increases, and decreases to such an extent that at the higher radio frequencies the flux set up may not be more than a very small per cent of the flux set up by the same current at 60-cycle frequency. As the force which actuates the finger of the meter varies with the square of the flux it is apparent that the meter reading will be practically zero at radio frequencies, no matter how much current is flowing through it.

This defect in the iron vane meter is due to the non-penetration of the flux at the high frequency; the error is present whether the meter is ammeter or a voltmeter.

In the case of an iron vane voltmeter the inductance of the coil which magnetizes the iron vanes, causes an additional error. A 100-volt range meter has about 800 ohms resistance and an inductance of 0.1 henry. The current which flows through such a meter (and it is this current, not the impressed voltage, which actuates the meter) varies with frequency, because of the effect of the coil reactance. Assuming one hundred volts impressed, the current which flows through the meter at different frequencies is calculated as follows:

Frequency	Resistance	Reactance	Impedance	Current	Fraction of true reading, per cent
25	800	15.7	800	0.125	100
60	800	37.7	800	0.125–	100
120	800	75.4	803	0.1243	99.5
500	800	314	858	0.1165	93.3
5,000	800	3,140	3,230	0.0309	24.7
100,000	800	62,800	62,800	0.00159	1.27
1,000,000	800	628,000	628,000	0.000159	0.127

In the above calculation it is assumed that the resistance is independent of the frequency; such is actually not the fact because of the "skin effect." This results in an increase in resistance, which is, however, negligible in the above calculation.

In the case of an iron vane ammeter there is no error corresponding to that calculated above for the voltmeter but, owing to the reactance of the ammeter, the impedance drop in the meter itself (which should be negligible) is hundreds, and even thousands, of volts at radio frequencies. Of course this means that the ammeter would either break down because of this excessive voltage strain on its insulation or would act as a choke coil, and so practically prevent the flow of radio frequency current in the circuit where it was used.

In the case of the dynamometer type meter there is no error due to the non-penetration of magnetic flux, but the effect of reactance is present as in the type just discussed. There is an additional effect in the dynamometer type which assumes importance at the higher frequencies. The two coils, one rotatable inside the other, are close together and so act as the two plates of a condenser. At high frequencies much of the current flowing between the two terminals of the meter does not flow through the two coils in series, as it should, but passes from one coil to the other across this condenser. This "condenser by-pass" effect also exists to some extent with the iron vane type of meter, passing from one layer of the coil to the other by means of the inter-layer capacity. The idea, as applying to a dynamometer type of meter, is illustrated in Fig. 2.

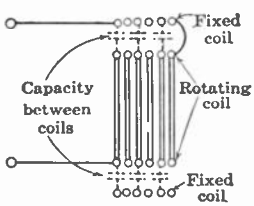


FIG. 2.

The ordinary wattmeter cannot be used in radio circuits at all. Its potential circuit is subject to the errors discussed above for the voltmeter; its current coil is subject to the same limitations as those discussed above for the ammeter; and in addition there is an error due to the fact that the current in the potential circuit is not in phase with the voltage in the potential circuit, but lags nearly 90 degrees behind the voltage at high frequencies. This error, caused by the reactance of the potential circuit of the wattmeter, is quite appreciable, even at 60 cycles, when the power factor of the measured circuit is low. This error will be discussed in detail in Exp. 1, where it must be taken into account to get accurate measurements of the resistance of a low power factor coil.

Amount of Power Used by Ordinary Meters.—The amount of power used by the meters discussed above depends somewhat upon the current and voltage range of the meter in question. In ranges less than one ampere the iron vane ammeter uses from 0.5 to 1.0 watt of power, at full scale reading; in ranges above one ampere the power is somewhat greater, say from 1.0 to 2.0 watts for meters of a few amperes range. Voltmeters of the iron vane type have a resistance of about 15 ohms per volt of scale range; thus a 30-volt meter would use about 2 watts of power at full range, and a 150-volt meter would use about 10 watts of power.

Rectifier Type of Alternating Current Voltmeters.—With the advent of the copper-copper oxide rectifier it became possible to build low range alternating current voltmeters, having a high resistance, a result impossible of attainment with any other type of rugged portable meter. The properties of this rectifier are examined in detail in Exp. 15, so they will not be discussed here. Fig. 3 shows how this rectifier is used, in combination with a sensitive c.c. ammeter, to make an alternating current voltmeter. Four small rectifying units are connected in a bridge scheme, poled as shown in Fig. 3. By inspection of this figure it will be seen that, for either polarity of voltage across the terminals, the current through the continuous current milliammeter A is in the same direction.

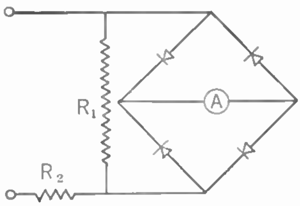


Fig. 3.

Although the scale of this meter is not quite uniform, as in a c.c. meter, it is much more so than any other type of alternating current meter. As commercially available, these meters have a resistance of about 2500 ohms per volt of scale. It is sometimes found advisable to put a high resistance R_1 in parallel with the bridge of rectifier units; a series resistance R_2 serves to increase the voltage range of the meter.

In one make of this copper oxide meter available on the market, it is stated by the manufacturer that the resistance is 2500 ohms per volt of range, and that the error is 10 per cent at 10,000 cycles. Actually a very peculiar error exists in the reading of this meter, in that it reads higher as the frequency is increased. And unfortunately its resistance, or rather its impedance, decreases rapidly as the frequency is increased. Thus a 5-volt meter shows an impedance of 13,000 ohms at 100 cycles, about 10,000 ohms at 500 cycles, and 300 ohms at 50,000 cycles.

A voltage of 5.00 volts was impressed at varying frequency, with the following unexpected result:

Frequency..	100	1 kc.	2 kc.	5 kc.	10 kc.	14 kc.	20 kc.	40 kc.	60 kc.	80 kc.
Reading....	5.00	5.08	5.22	6.13	7.80	8.77	9.52	10.2	10.3	10.3

There is a resistance unit of about 10,000 ohms used for R_2 (Fig. 3), and apparently the internal capacity of this resistance unit acts as a partial short circuit to the resistance at higher frequencies, thus impressing a higher proportion of the impressed voltage on the rectifier units and so giving a higher reading.

In another meter of this type available on the market, the reading was lower at the higher frequencies; it is possible, however, so to design and build these rectifier meters that they read correctly (to within a few per cent) for frequencies as high as 100 kc.

The resistance of this copper oxide type of meter is different for a high-scale reading than for a lower one; this effect becomes more noticeable the higher the frequency.

Hot Wire Meters.—Various grades of hot wire ammeters are available, ranging in price from $5 to $100, according to the type of construction used. This type of meter is used quite extensively for high frequency measurements, but it uses a considerable amount of power, is very easy to burn out, and is not at all permanent in its calibration.

A thin platinum (or other suitable metal) wire or strip carries the current to be measured; its expansion due to the heat generated, operates through a multiplying scheme to actuate the pointer. A simple illustration of the arrangement is given in Fig. 4. In common with all meters of this type, which give a response varying with the square of the current, its scale is very much crowded in the lower range. A 1.0 ampere meter, for example, is not even calibrated below about 0.2 ampere, so that the low readings are very inaccurate, especially when the "creeping zero" is taken into account. On most meters of this class it is necessary continually to readjust the zero setting, as the meter is used.

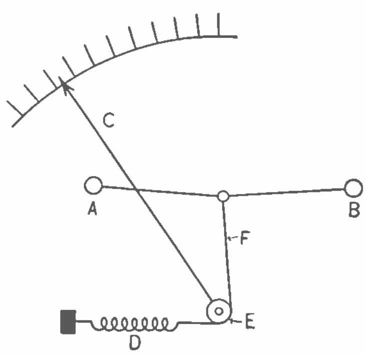


Fig. 4.

There is some variation, between different makes, as to the amount of power used, but a typical line of commercial meters, costing about $20 each, has resistance and power loss values as given on page 7.

The low range meters, using the smaller amount of power, are necessarily more delicate in construction than the higher range ones.

One can realize easily that it is impossible to build hot wire meters of very low current range. The heating falls off with the square of the current, so that even if very fine wire is used the heat available is

Range, amperes	Resistance, ohms	Power for full scale, watts
0.1	31.0	0.31
0.25	7.0	0.44
0.50	3.5	0.87
1.0	1.2	1.2
2.0	0.5	2.0
5.0	0.15	3.7

vanishingly small for low currents. The most sensitive meter of this type available on the market has a resistance of 7 ohms and requires .05 ampere to deflect the finger full scale. The scale is very short, and the frictional errors in the meter are large.

It will be evident that hot wire voltmeters are not very useful in the average radio experiment, because of the very considerable amount of power required. Such a meter uses about 0.15 ampere for full scale deflection, so that a 150-volt range meter requires over 20 watts of power. This is more power than is generally available, unless the experiment has to do with a high-powered transmitting set.

Thermocouple Meters.—The radio laboratory uses this type of meter, both for voltmeter and ammeter, more frequently than any other.

In the most generally employed form a copper-constantan junction is heated by a fine wire through which flows the current to be measured. The couple is straddled over the wire (sometimes welded to it) so that the couple hot junction is in direct contact with the current-heated wire.

With temperatures safely used in such meters a single junction will generate a few millivolts. The couple is connected to a sensitive continuous current millivoltmeter, this generally having a uniformly graduated scale.

As the voltage generated by the thermocouple is directly proportional to temperature, and the temperature of the heated wire is proportional to the square of the alternating current flowing in the heater, the reading of the millivoltmeter is proportional to the square of the current through the heater wire.

Thus, suppose the scale has 100 divisions, and that full scale reading in the millivoltmeter is obtained with a current of 10 milliamperes in the heater. The millivoltmeter will read 25 divisions when 5 milliamperes is flowing and 1 division when 1 milliampere is flowing through the heater.

In order to make these couples as sensitive as possible (greatest temperature for a given power), the heater wire must be of the proper length, it should be smooth and bright, the thermal junction should be small, and yet of low resistance, and the millivoltmeter used should be sensitive and have a resistance approximately equal to that of the couple.

To conserve the heat generated in the heater wire the whole junction and heater are mounted in a small glass bulb from which the air has been evacuated. This removes the possibility of the surrounding air cooling the heater wire. The heat-radiating efficiency of the heater wire, and its supports, should be low so that a given heat may result in a maximum temperature. Fig. 5 shows the normal construction.

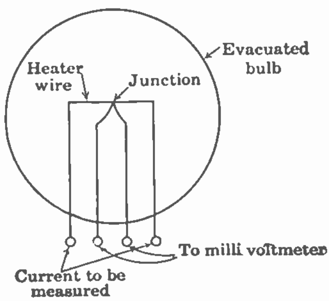


Fig. 5.

The couple is really a small generator, and the student of electrical machinery knows that the load circuit should have the same resistance as the armature of the generator itself, if maximum power is to be delivered by the generator to the load. This is the reason for winding the coil of the millivoltmeter with such size of wire as gives a resistance equal to that of the thermocouple itself.

It is to be pointed out that nothing is to be gained by winding the millivoltmeter coil with german silver wire, or filling only a part of the available winding space, in order to equalize the resistances of coil and couple. All the available winding space must be filled with wire of the lowest specific resistance possible. The size of wire should then be selected to give the proper resistance; the smaller size wire of course permits a greater number of turns in the coil.

The consideration of heat radiation from the wire itself, of heat loss by conduction to the heavy wire supports of the heater, heat delivered to the couple itself, and of the desire to get a maximum temperature at the couple, require a heater wire about 1 cm. long and of a few mils diameter, or less, depending upon the current to be passed through it. Low-resistance heaters are metallic and high-resistance heaters are of carbon. The resistance and current capacity of the ordinary vacuum couples are about as given on page 9.

For current ranges greater than those given in this table it is not worth while to employ a vacuum; the couple in air uses somewhat more power but is quicker in action than the vacuum couple.

Normal current....	0.002 amp.	0.005	0.01	0.025	0 05	0.10	0.30
Maximum current..	0.004 amp.	0.010	0.015	0.035	0.065	0.13	0.35
Heater resistance...	750 ohms	150	30	10	5	1.5	1.0

The couples themselves are the same whatever the heater resistance; these couples have a resistance of about 10 ohms and generate about 5 millivolts with normal heater current. If a 10-ohm millivoltmeter, of 2.5 millivolts full range, is then connected to these couples, the meter will read full scale when normal current is flowing through the heater.

It is possible to get a portable millivoltmeter of the suspension type, having 10 ohms resistance, which gives full scale reading with about 0.25 millivolt. Using this meter with the 750-ohm heater couple, full scale reading will be obtained with about 600 microamperes in the heater. If the scale of the millivoltmeter has 100 divisions it will require 60 microamperes in the heater to give one division deflection on the millivoltmeter.

If a wall type suspension millivoltmeter is used it is possible to get a just readable deflection with from 10 to 20 microamperes in the heater. This is the smallest radio frequency current that can be directly measured.

These thermocouple ammeters are not subject to appreciable frequency errors, within the frequency range used in the average experiment.

It is evident that the thermocouple millivoltmeter combination can be used as a voltmeter. Thus, the first one listed above required 0.002 ampere in a 750-ohm heater to give full scale deflection. But to force 0.002 ampere through 750 ohms requires 1.5 volts, hence the combination may be used as a voltmeter, with a 1.5-volt range. Whether used as ammeter or voltmeter it is to be remembered that the reading *varies as the square* of the quantity being measured.

The vacuum couple itself costs about $12, and a suitable millivoltmeter costs about $50. When an overload occurs the couple-heater combination burns out and must be replaced by a new one; the millivoltmeter is seldom injured, however.

Efficiency of Thermocouple.—Evidently this combination serves to change energy from that in an alternating current circuit to energy of a continuous current circuit. It is interesting to calculate the efficiency of conversion.

Let us calculate the amount of alternating current power used in typical heaters, and the amount of continuous current power generated by the couple. Whatever the heater resistance the couple has 10 ohms resistance and the millivoltmeter has 10 ohms resistance, and the voltage generated by the couple is about 5 millivolts, so that the power generated for full scale meter deflection is $\frac{.005^2}{20}$, or (approximately) 1 microwatt, equals 10^{-6} watts.

The power used in the heaters is given below:

Resistance	750 ohms	150	30	10	5	1
Normal current	0.002 amp.	0.005	0.01	0.025	0.05	0.30
Power used	0.003 watt	0.0037	0.003	0.00625	0.0125	0.09
Power generated	10^{-6} watts	10^{-6}	10^{-6}	10^{-6}	10^{-6}	10^{-6}
Per cent efficiency of conversion	0.03	0.026	0.03	0.016	0.008	0.001

Moreover, as the c.c. power varies as E^2_{thermal} and as E varies as I^2_{heater} it follows that the efficiency of conversion, for a given heater, varies with the square of the current through the heater. The efficiencies given above are for normal current, hence the actual efficiency of a thermocouple in use is this much or less, falling off rapidly with smaller currents in the heater. For example, with only half the rated normal current in the heater the efficiency is only one-quarter as much as the value given in the table.

Both the hot wire meter and the heater of the thermocouple might be expected to change their resistances according to the amount of current passing through them, because of the large temperature changes taking place.

The hot wire ammeter and lower resistance heaters increase their resistances with current, the increase sometimes being 10 per cent from zero to full scale reading. The high resistance heaters, however, are made of carbon and have negative temperature coefficients, decreasing their resistances from 5 to 10 per cent at full scale reading. Actual data on this feature are obtained in one of the experiments.

What Meters Read on Complex Waves.—An alternating current meter gives a force which varies with the square of the current flowing through it; this is different from the common form of continuous current meter (the D'Arsonval type) in which the force varies with the first power of the current.

The reading of an a.c. meter is proportional to the square root of the mean square of the instantaneous values of the current, taken over one cycle. In case the current through the meter is a simple sine wave this is equal to $1/\sqrt{2}$ of the maximum value.

In case the wave is complex in form (but periodic) it is made up of two or more sine waves, all having a definite frequency fixed by the periodicity of the complex wave. One of the component waves (generally the principal one) has the same frequency as the complex wave itself. All the other component waves are sine waves having frequencies which are multiples of this fundamental frequency.

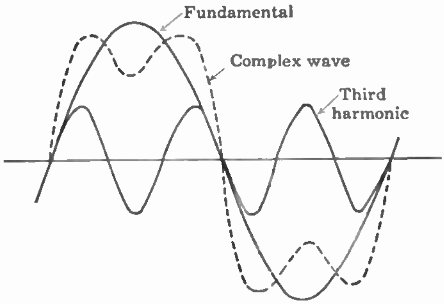


Fig. 6.

In Fig. 6 is shown a complex wave made up of a fundamental sine wave of amplitude 5 amperes, and a third harmonic current of amplitude 2 amperes. In Fig. 7 is shown another complex wave, made up of the same two components (5 amperes fundamental, 2 amperes third harmonic), but the phase of the third harmonic has been shifted with respect to that of the fundamental. In general, a complex wave is made up of a great many components, the magnitude of the component decreasing as its frequency is higher

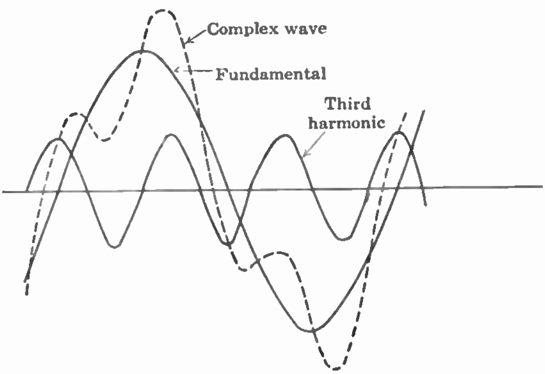


Fig. 7.

If a wave has sharp corners on it, as indicated in Fig. 8, there must be a great many harmonics in its composition, the sharper the corner the more harmonics are there present. To represent the wave shown in Fig. 8 with a reasonable accuracy, probably twenty or more harmonics must be used, and even then the peaks would be slightly rounded instead of true points.

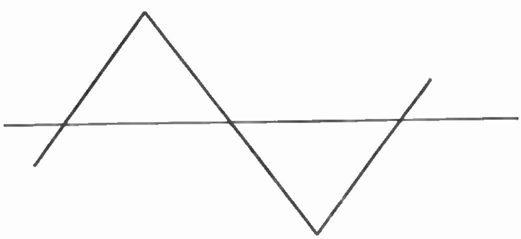


Fig. 8.

When the positive and negative loops of the wave have the same form there are no even harmonics present; even harmonics make the two alternations dissimilar. In Figs. 6 and 7 only odd harmonics are present; in Fig. 9 is shown a wave having both even and odd harmonics.

FIG. 9.

In many radio circuits a pulsating current exists; it is unidirectional, but has large fluctuations of amplitude. Such a wave is shown in Fig. 10. In such a wave there is a continuous current component, and a whole series of alternating current components. The c.c. component is the *average* of the pulsating current; this average value is shown in the dashed line in Fig. 10. The form of the actual current, with this dashed line as a zero axis, is the alternating current part of the pulsating wave. It is generally made up of many sine wave components.

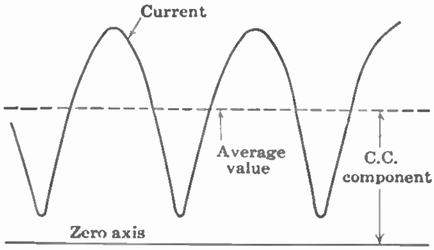


FIG. 10.

An alternating current meter carrying a complex current reads the square root of the sum of the squares of the effective values of all the component waves. Thus, suppose a wave consists of a fundamental wave of 10 amperes maximum value, and a third harmonic of 5 ampere maximum value. The effective value of the fundamental is 7.07 amperes, and of the third harmonic the effective value is 3.53 amperes. The meter will then read $\sqrt{7.07^2 + 3.53^2} = 7.9$ amperes. Suppose a pulsating wave similar to that of Fig. 10 is given by the equation

$$i = 15 + 8 \sin wt + 7 \sin 2\,wt + 5 \sin 3\,wt + 3 \sin 4\,wt \quad . \quad (1)$$

The effective values of the sine components are equal to $1/\sqrt{2}$ of their amplitudes given in this equation; the effective value of the continuous current component is the same as its actual value. The ammeter through which this current is flowing would then read

$$I = \sqrt{15^2 + \frac{8^2}{2} + \frac{7^2}{2} + \frac{5^2}{2} + \frac{3^2}{2}} = 19.35 \text{ amperes} \quad . \quad . \quad (2)$$

If a continuous current meter were in series with the alternating current meter it would read the average value of the pulsating current, namely, 15 amperes.

If a pulsating current is known to consist of a simple sine wave fluctuation, the amount of fluctuation can be easily obtained by the readings of an a.c. ammeter and a c.c. ammeter through which the current is flowing. Such a wave is shown in Fig. 11, and we suppose the c.c. ammeter reads 10 amperes and the a.c. ammeter reads 11 amperes.

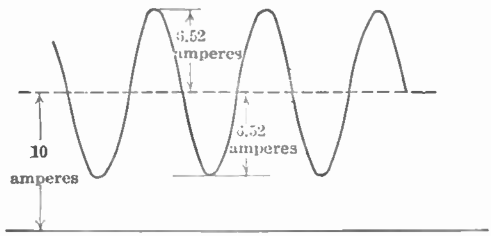


FIG. 11.

Remembering that this a.c. meter reads the square root of the sums of the squares of the effective values of the components, we find that the effective value of the sine wave component is equal to

$$\sqrt{11^2 - 10^2} = \sqrt{21} = 4.6 \text{ amperes.}$$

The maximum value of the sine wave is then

$$4.6 \times \sqrt{2} = 6.52 \text{ amperes.}$$

The wave in question then has an average value of 10 amperes with a fluctuation of 6.52 amperes both above and below this average value. These values are indicated in Fig. 11.

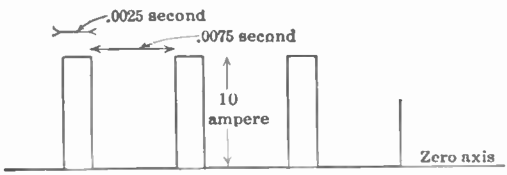


FIG. 12.

Suppose a rectangular pulsating wave such as shown in Fig. 12 is flowing through a c.c. ammeter and a.c. ammeter in series. What will each read?

The current of 10 amperes is flowing for just one quarter of the time, and the rest of the time the current is zero. The average current is evidently 2.5 amperes, and this is what the c.c. meter will read. The square root of the average square is obtained by noting that the squared value (100) lasts only one-quarter of the time, hence the average square is 25. The square root of this is 5. Hence the a.c. ammeter through which this current is flowing would indicate 5 amperes. That is the alternating current meter would read just twice as much as the c.c. ammeter—and both of them are right.

This rectangular pulse current is somewhat the shape of charging current supplied to a storage battery by a "half wave" rectifying tube. The value of the current, in so far as charging the battery is concerned, is 2.5 amperes, but the value of this same current, in so far as heating the battery by I^2R loss is concerned, is 5 amperes!

Feasibility of Shunting an Ammeter.—In general, it is not good practice to use shunts on ammeters operating in radio frequency circuits, because a shunt which changes the meter reading, say, 5 to 1 at one frequency, will not have the same multiplying factor for another frequency. However, *for a given frequency*, it is not only feasible but is generally advisable in the laboratory to get into the habit of shunting meters. Such practice cuts down the necessary number of meters.

A decade resistance box makes a convenient shunt; the box should have a total resistance about the same as the resistance of the ammeter or thermocouple with which it is to be used. The circuit is adjusted so that the ammeter (hot wire or thermocouple) reads well up on its scale; the resistance box is then connected in parallel with it (using leads as short as possible), and by proper adjustment of its resistance the reading of the meter is reduced say to one-fifth. The circuit may then be left in this adjustment and all readings of the meter multiplied by five.

If the frequency is altered by an appreciable amount the multiplying factor for the shunt must be redetermined, as it is likely to change considerably if the frequency is changed very much.

Wave Meters.—This type of meter, which finds application only in the radio field, should really be called a frequency meter. It is nothing but a calibrated resonance circuit, including generally a resonance-indicating device.

A low-resistance coil (generally one of a set of coils) and a continuously variable condenser of rugged construction, are put in series with a suitably low-resistance thermocouple or hot wire ammeter. Such a circuit, if excited by a radio frequency voltage, will show a maximum current when its total reactance is zero, that is, when the coil reactance is equal to the condenser reactance.

For a given setting of the condenser, and given coil, only one definite frequency will bring this about, hence this point on the condenser scale may be calibrated at that frequency. The various condenser settings will correspond to different resonant frequencies, and these frequencies, instead of capacity values, may be marked on the condenser scale. Or this may be calibrated in divisions only, and suitable curves be furnished with the meter showing the resonant frequency for each condenser setting. By using a set of half a dozen properly proportioned coils, the same condenser may be used to cover a wide frequency range, the condenser generally being good for a three to one frequency range with each coil.

For convenience the various coils overlap one another slightly in frequency range; thus the first coil might be good for a range of 100 kc. to 300 kc., the second for 275 kc. to 825 kc., the third 775 kc. to 2300 kc., etc. Ordinarily one coil is given the proper inductance to cover the broadcast range of frequencies, 500 kc. to 1500 kc.

Wave meters designed to cover the highest frequency range are extremely simple; a small variable condenser of perhaps two moving plates and three stationary plates, combined with a coil consisting of say one turn of wire of No. 10 wire, diameter of the one turn being about four inches, covers the range 20,000 kc. to 50,000 kc. No current-indicating device is used to indicate resonance; the skilled observer can tell when the wave meter is in resonance with the oscillating circuit of his vacuum tube by the indications of the meters in the tube circuit itself.

As the wave length of a radio wave and the frequency of the current setting up the waves are related by the simple formula

$$\lambda_{\text{meters}} = \frac{300{,}000{,}000}{f(\text{cycles per second})} \quad . \; . \; . \; . \; . \; . \quad (3)$$

wave meters can readily be calibrated in either frequency or wave length. At present they are generally calibrated in terms of wave length.

As stated before, the ordinary type of wave meter is equipped with a thermocouple ammeter to indicate resonance, but there are many times when it is desired to measure wave length when the power available is too small to give any indication on this meter. In such case a telephone, in series with a crystal rectifier, is always used. There are several different schemes used for exciting the phones and crystal from the wave meter, two of them being shown in Fig. 13. That of *a* is the normal connection, giving a much greater response in the phones than the arrangement of *b*, which is called the unilateral connection. However, connection *a* seriously increases the decrement of the wave meter, and has a very appreciable effect on the calibration of the wave meter, especially for the smaller values of the wave meter condenser. For such a connection two calibration curves should be furnished, one when the ammeter is used as resonance indicator and one when crystal and phones are used. Generally the ammeter stays in circuit even when phones are used as indicator.

The connection of *b*, Fig. 13, has but little effect on either the decrement or calibration of the wave meter; it does, however, require that considerably more power be supplied to the wave meter than is required for the other scheme.

A good wave meter has a decrement not greater than 0.02; this means that the reactance of the coil (or condenser) is more than 100 times as much as the total resistance in the wave meter circuit. The effect of crystal and phones on the decrement of such a meter is large.

Suppose that the wave meter condenser is set at the value of 0.001 μf (microfarad) when measuring a wave length of 1000 meters (300 kc.), and that the effective a.c. resistance of the phone and rectifier circuit across this condenser is 50,000 ohms. The reactance of the condenser is practically 500 ohms, so that the equivalent series resistance of this shunt circuit is $500^2/50{,}000$, or 5 ohms.* Now, if the reactance of the coil of the wave meter is 150 times the normal resistance of the wave meter this must be 500/150, or 3.3 ohms. Hence, without the phones and rectifier across the wave meter condenser the wave meter resistance is 3.3 ohms; when the phones and crystal are used for finding resonance the resistance (series) of the wave meter is 8.3 ohms. This means that the use of phones and crystal detector makes the resonance curve of the meter more than twice as wide as it is when using the ammeter above for indicating resonance.

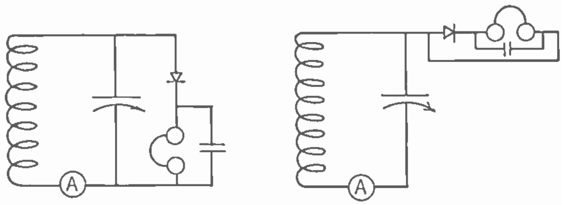


FIG. 13.

It is evident that in wave meters using tuning condensers of low capacity the effect discussed above is even more pronounced; in such meters the arrangement shown in circuit *b* of Fig. 13 must be used if the resonance curve is to be kept sufficiently narrow to permit accurate measurement of wave length.

In addition to increasing the decrement of the wave meter, as shown above, the connection of the rectifier-phone circuit across the tuning condenser must necessarily increase the capacity in the wave meter circuit, and hence result in a lower resonance frequency. As the added capacity is not large, perhaps 25 or 50 $\mu\mu f$, the effect on the resonance frequency will not be large unless the capacity of the wave meter condenser itself is small.

Suppose that at its lowest useful setting the wave meter condenser has a capacity of 100 $\mu\mu f$. The addition of 25 $\mu\mu f$ would change the

* See Principles of Radio Communication, 2nd ed., p. 222.

resonance frequency in the ratio of $\sqrt{\frac{1.25}{1.00}}$, or 1.12, that is, the wave meter calibration at this point would show a frequency calibration 12 per cent lower with crystal and phones than without them. But when the wave meter condenser is being used at its maximum capacity say 3000 $\mu\mu f$, then the addition of the 25 $\mu\mu f$ will result in a decrease in frequency of less than $\frac{1}{2}$ per cent. In Fig. 14 are shown the two calibration curves; the dashed line applies when using the crystal and phones, and the full line when the ammeter only is being used as resonance indicator.

Most wave meters are arranged so that when properly connected to a small buzzer and battery the wave meter becomes a generator of damped waves of known wave length. Binding posts are put on the

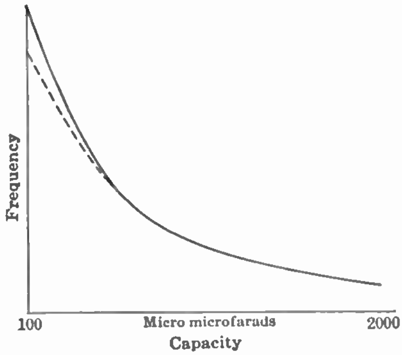


Fig. 14.

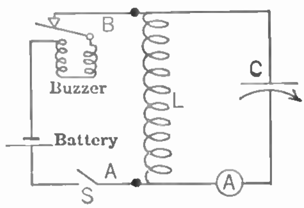


Fig. 15.

meter for making the connection shown in Fig. 15. When the switch S is closed, a current I flows through the coil of the wave meter. When the current, rising on the logarithmic curve

$$i = I_{\max}(1 - \epsilon^{-at}) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

becomes large enough, the buzzer armature pulls away from the contact point, thus opening the buzzer circuit. At this instant there is an amount of energy equal to $\frac{Li^2}{2}$ stored in the wave meter coil, as well as a very small amount in the wave meter condenser. This energy starts oscillations in the L-C circuit at the natural frequency of this circuit. They are damped oscillations, dying away at a rate fixed by the decrement of the wave meter circuit. These damped waves have disappeared completely before the buzzer armature, by moving back to the contact screw, again starts the operation. Inspection of Fig. 15 shows that in this use of the wave meter there is an additional capacity connected across the tuning condenser, namely, the capacity between wires *A* and *B*. So in this use of the wave meter a correction must be made to the calibration curve, just as was done when using crystal and phones.

Rheostats.—Of course variable resistances are always required in any engineering laboratory, whatever the frequency of current being used. The type of rheostat frequently used for laboratory work in power frequency circuits is of practically no use in the radio laboratory. An iron pipe covered with insulation of some sort is wound with high-resistance wire, and a slider moving the length of this wire-wound pipe permits the variation of resistance from a maximum to zero.

When used in a radio circuit such a rheostat is subject to many errors. The wire used is frequently rather heavy, and owing to skin effect, at radio frequencies may have many times as much resistance as at 60 cycles.

Also the iron pipe constitutes a short-circuited secondary winding, closely coupled to the real rheostat winding, and this short-circuited secondary introduces an added resistance, of unknown amount, in the rheostat circuit.

Next to be considered is the fact that at the higher radio frequencies the current does not traverse the length of the rheostat wire, but goes by a capacity path from the resistance wire into the iron pipe, along the iron pipe and again by capacity path to the other end of the resistance wire. This idea is illustrated in Fig. 16; the two condensers marked C_1 and C_2 represent the capacity between the end sections of the winding and the iron pipe. As the winding is separated from the iron pipe by only a thin coat of enamel, this capacity is much larger than one would think. In Fig. 17 is shown the experimentally determined impedance of one of these enameled iron pipe rheostats; its shape is dependent upon the combined effects of resistance, eddy currents, inductance, and capacity.

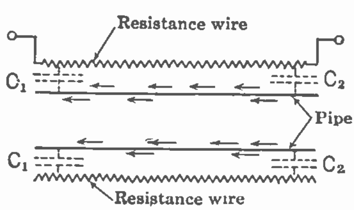


FIG. 16.

This capacity by-pass effect is of appreciable magnitude only in high-resistance rheostats, or at very high frequencies. If the wire of the rheostat is of low resistance, only a negligible portion of the current will flow through this capacity by-pass. Furthermore, these rheostats which act like essentially non-reactive resistances at 60 cycles have a reactance at radio frequencies which may be many times as great as their resistance.

There are, however, available for the radio laboratory special pipe-wound rheostats which function very well. They are wound on a porcelain pipe, and have bifilar windings. The resistance wire is made up of two wires, one around the pipe in one direction and the other in the opposite direction. These two wires cross each other twice per turn as they progress down the length of the rheostat. Such a rheostat is practically non-inductive, the magnetomotive force set up by the current in one wire being practically neutralized by that set up by the other, flowing in the opposite direction.

Being inductanceless there is a correspondingly small skin effect,

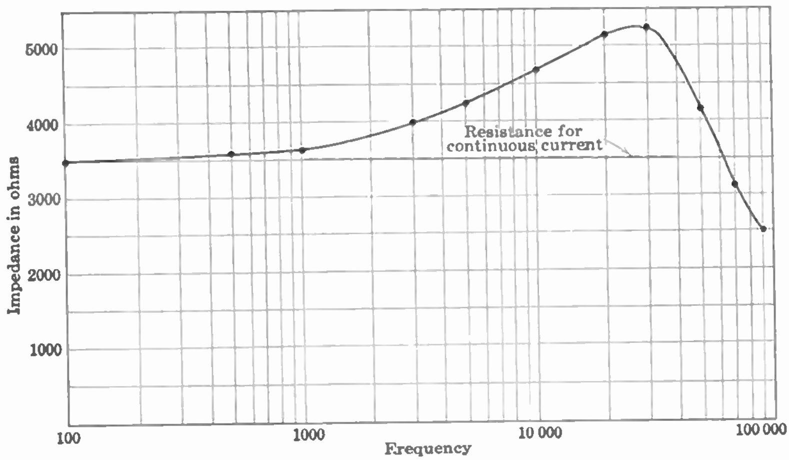


FIG. 17.

so that the resistance at radio frequencies is not greatly increased over the continuous current value.

For use in circuits carrying comparatively small currents, resistance boxes are used instead of rheostats. At radio frequencies these may be very inaccurate, even though they are the so-called "standard resistances" of the instrument laboratory. Correct for continuous current, or alternating currents of low frequencies, these may be inaccurate by hundreds of per cent when used in radio frequency circuit. As they are generally made, the various units of a resistance box may show a high capacity or a high inductance, depending upon how the resistance unit is wound (bifilar or not), the resistance value, and the frequency used.

Special resistance boxes must be used in radio frequency circuits. The resistance wire is wound on very thin mica sheet, many times bifilar, and so are nearly free from skin effect, and have very small inductive or capacitive reactance.

Even when these specially built resistance boxes are used (the resistances being taken at their specified values) the inductive reactance can not be neglected when using the low resistance units. Thus, suppose the one ohm unit has an inductance, in itself and its connecting wires, of one microhenry. At a frequency of 1000 kilocycles the reactance is more than six times the resistance. And even though the resistance units are correct, at their rated values, it will generally be found that the ten 0.1-ohm units are not exactly replaceable by one 1.0-ohm unit. The wires connecting the different units to the binding posts of the box have appreciable inductance, and as these wires have different lengths the reactance of the ten 0.1-ohm units will quite likely be different from that of the single 1.0-ohm unit.

This error, it is to be noticed, is not due to the resistance being other than its specified value, but to the presence of reactance. In case the resistance is being used in a tuned circuit the reactance of the resistance units is easily compensated by a slight readjustment of the tuning condenser.

When the resistance box is being used as a voltage divider, however, the error due to reactance is by no means negligible. Thus, it might be assumed that a current of 0.001 ampere (read on a thermocouple) flowing through the ten 0.1-ohm units of the resistance would give a drop of 1 millivolt. Actually, owing to the reactance, the drop might possibly be 5 millivolts if the frequency of the current was a million cycles, and correspondingly greater for higher frequencies.

This error, due to unavoidable reactance, becomes negligible in the units of higher resistance; it is only in the ohm and fractional ohm units that the error is appreciable.

Batteries.—Whereas most radio sets built and used today are independent of batteries for their operation, it is still convenient in the radio laboratory to use batteries as a source of filament power, as well as for grid and plate circuits.

For filament circuit supply, portable storage batteries of the Edison type are the most satisfactory in ordinary laboratory use. They are not as efficient as the lead-acid cell but they stand abuse much better than a lead cell. Thus, if left for a long time with insufficient electrolyte, perhaps discharged, they will be as good as ever after having the electrolyte replenished, and being charged and discharged a few times; a lead cell would be permanently injured by similar treatment.

Cells of about 50-ampere-hour capacity are right for average laboratory use; they are conveniently put up in carrying racks which hold five cells.

For plate circuit supply the radio laboratory should have a 500-volt storage battery of the lead-acid type with plenty of taps at various stations around the laboratory. There should be taps at 22, 45 67 90, 130, 200, 300, 400, and 500 volts.

If the battery is of sufficient ampere-hour capacity (50-ampere-hour cells are advisable) high-resistance potentiometers may be connected across the various parts of the battery without doing any harm. In this way any voltage up to the maximum of the battery is readily obtainable; for many tests, of course, the tap voltages themselves will serve so that the potentiometer connection is unnecessary.

In addition to the main high-voltage battery there should be blocks of dry cells for plate and grid voltage. It does not pay to get the small-sized dry cells as they disintegrate quickly, even when not being used. Those of about one-ampere-hour capacity are suitable for general use.

The student should be cautioned not to use a potentiometer scheme for getting variable voltage from these cells. Unless a very high potentiometer, is used, the cells may be completely used up in a few hours, not by the current taken by the tubes or other apparatus under test but by the current taken by the potentiometer itself.

A few boxes filled with individual (or three in a block) dry cells are convenient. Ten of the three cell blocks, packed into an open wood box, connected in series, give 45 volts. The terminals of each cell being available, it is possible to get voltage in 1.5-volt steps up to the full 45 volts.

A few such boxes of individually available cells, to operate in series with the 45-volt blocks, permit the experimenter to get any voltage he may wish without the necessity of putting a potentiometer across the dry cells.

Coils Suitable for Low Frequency Experiments.—In order that the 60-cycle tests may give results reasonably comparable with those obtained at radio frequencies, the coils must have a high ratio of reactance to resistance, that is, a low power factor. In ordinary 60-cycle apparatus a power factor of 0.2, or even 0.3, is considered very low but this is much too high for apparatus designed to simulate that used in radio circuits.

To get a low power factor a large amount of copper must be used, and the copper must be properly made up into a coil. In Fig. 18 is shown a 75-lb. coil having 10 ohms resistance and 0.6 henry inductance. It is wound with cable consisting of 7 strands of No. 20 enamel-covered wire. Its resistance at 100 cycles is practically the same as its continuous current resistance. There are 30 layers of 50 turns per layer. The winding is opened at the fifteenth layer; it has four ends brought out to make two coils each having 15 layers. The inner half of the coil has 0.14 henry and the outer has 0.2 henry of inductance; the mutual induction between the two sections is 0.13 henry.

At 100 cycles the complete coil (both sections in series) shows a reactance 35 times as much as its resistance, that is, a power factor of 0.03. If a lower power factor than this is desired, a larger coil (that is greater weight of copper) must be used. Such an air core coil has an inductance nearly independent of frequency, up to two or three hundred cycles, and its inductance is independent of current value.

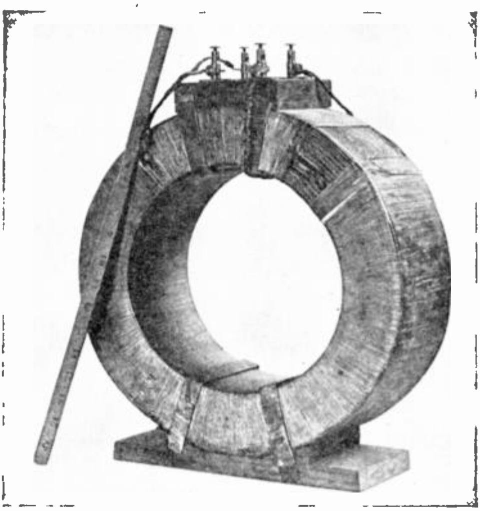


FIG. 18.

If the same size coil is wound of very fine wire, such as No. 36, it might be thought that a higher ratio of reactance to resistance would be obtained, but this is not so. The same weight of copper, wound into the same size of coil, gives the same ratio of reactance to resistance *no matter what size of wire is used.*

It might be thought that an iron core coil would be more suitable for getting a low power factor than an air core coil, but such is not the fact. Even with a well-laminated core, such as is used in transformers, the reactance is practically never ten times as much as the *effective* resistance, and generally only about five times as much.

Iron core coils, of course, are generally undesirable in that their inductance varies greatly with current, and appreciably with frequency, and they also distort the form of current in the circuit.

When inductance is defined from the standpoint of "interlinkages per ampere" the field coil of a large generator has a very high inductance, and if this inductance is used, in combination with the c.c. resistance of the field winding, a very large ratio of inductance to resistance is obtained. If this inductance to resistance ratio is used in calculating what the power factor will be on a 60-cycle line a remarkably low value will be obtained, leading one to believe that the reactance will be perhaps fifty times as much as the resistance. When the field coil is actually tested on an a.c. circuit, however, it will be found that the power factor is from 0.3 to perhaps 0.6, which means that the reactance is not more than twice as much as the resistance.

This great difference between the actual value, and the one expected, is due to the fact that the *effective* resistance (the resistance which is necessarily considered in an a.c. circuit) is many times as much as the c.c. resistance of the field coil, and the inductance in the a.c. test (change in interlinkages per unit change of current) is much less than the value obtained by the c.c. test.

If coils are to be used, in combination with condensers, in making series resonance tests, let us say, iron core coils are entirely out of the question. A coil having a closed iron magnetic circuit gives extraordinary behavior in such a circuit, and the relations obtained will completely hide the ordinary resonance phenomena the test is designed to show. If a study of parallel resonance is being made the iron core coil again gives very unsatisfactory results, owing primarily to the very distorted form of current it draws from the line.

In general, iron core inductances should never be used in a radio apparatus circuit, except where a large reactance is required, with a comparatively small cheap coil, and where the high effective a.c. resistance is not detrimental.

In circuits carrying radio frequency currents, iron core coils are entirely out of the question; the reactance may actually be less than the resistance if the frequency is very high. For frequencies up to about 100 kc., multilayer, bank-wound air core coils are best, and for higher frequencies single-layer solenoids are best. At very high frequencies it pays to actually space the turns of the single layer solenoid, one from the other; this spreading of the winding decreases the inductance of the coil somewhat but it decreases the resistance still more, so that the reactance to resistance ratio is greater than if the same number of turns are used, wound close to one another. For frequencies measured in tens of megacycles the coils should be wound of about No. 10 solid copper wire, and the space between consecutive turns should be about equal to the diameter of the wire itself.

Condensers.—The condensers most suitable for experiments at power frequencies are of the impregnated paper type; a carefully selected paper about one mil thick is used between the thin plates of tinfoil or aluminum foil. During manufacture this paper is freed from all moisture and air, and is impregnated with oil or paraffin wax. The single layer of paper is sufficient for condensers used on hundred-volt circuits; when higher voltages are to be impressed several layers of paper are used.

Such condensers are procurable in units of one or two microfarads at a cost of from $.25 to $2 or more per microfarad, the price going up as the voltage rating goes up.

A convenient method of arranging these condensers for laboratory use is to mount them in shallow wooden trays, in three or four rows of 10 microfarads each. One terminal of all condensers is connected together and to one terminal post of the tray. Two or three of the rows have their other terminals connected together, and to the other terminal post of the tray, there being a small single pole switch in each of these row connections. The last row has the free terminals of four condensers connected together, and this connection, through a small single pole switch, goes to the second terminal of the tray. Similarly for 3, 2, and 1 microfarads of this last row. By this scheme all of the thirty or forty microfarads may be readily used or any number of microfarads between one and the total number in the tray.

The voltage rating of a condenser is nearly always given in terms of the continuous voltage it will stand without puncturing. Only a fraction of this amount must be used when the condensers are operating on an alternating current circuit, for two reasons.

The peak voltage is what punctures a dielectric, and the peak value of a sine wave is 41 per cent greater than the voltage read in a voltmeter. Thus a 110-volt circuit impresses on a condenser a voltage as high as 155, twice per cycle.

There are losses occurring in the dielectric on an a.c. test which do not exist in a c.c. test, so the dielectric warms up more. And as a dielectric like wax warms up, its ability to withstand puncture decreases very rapidly.

Because of the two facts mentioned above, a condenser should not be used on an a.c. circuit of voltage rating greater than about 60 per cent of the c.c. rating of the condenser, and this factor decreases as the frequency of the alternating voltage increases. Thus, a condenser with a c.c. breakdown voltage of 500 should not be used in an a.c. circuit of greater than 300 volts. The condenser might stand, momentarily, considerably more than this, but if more than 300 volts is continuously applied it is very likely to puncture.

The losses in good condensers of the wax or oil-impregnated type are very small, varying of course with the quality of the dielectric. The current will generally lead the voltage by more than 89°, which means that the power factor of the condensers is 1 per cent or less. This means that the *equivalent series resistance* of the condenser is not more than 1 per cent of its reactance. This series resistance is entirely different from the *insulation resistance* of the condenser. This should be above 10 megohms per microfarad and in good condensers may be as high as 1000 megohms or more per microfarad.

The equivalent series resistance of a good condenser is generally sufficiently small in a 60-cycle circuit so that when it is used in series with coils no great error is involved in assuming that all of the circuit resistance resides in the coil. However, if it is desired to allow for the resistance of the condenser it may be assumed as 1 per cent of the reactance of the condenser for paraffin paper, about $\frac{1}{2}$ per cent of the reactance for an oil-impregnated paper, and about $\frac{1}{5}$ per cent for a condenser using mica for its dielectric.

In many pieces of apparatus used with radio receivers and transmitters it is desired to obtain much more capacity in a given volume than wax or oil types furnish. Frequently the electrolytic condenser is used in such cases. Oxidized aluminum plates in a boric acid solution constitute the working parts of these condensers. The thin oxide film forms the dielectric, the aluminum plate and the solution being the two "plates" of the condenser. Such condensers have extraordinarily large capacity values, because of the small thickness of the dielectric. In the space required for 2 microfarads of paraffin wax condenser we may put 50 to 100 microfarads of electrolytic condenser.

Electrolytic condensers require a continuous voltage, of proper polarity, for maintaining the oxide film, otherwise the film disintegrates and the condenser, as such, disappears.

The boric acid solution (sometimes mixed with glycerin when danger of freezing exists) is in a glass or copper can; in the latter case the copper may form one side of the condenser. The oxidized aluminum is generally in the form of thin foil, rolled or corrugated so that a large area is in contact with the electrolyte.

This aluminum electrode must be connected to the positive terminal of the required polarizing source of continuous current; if the opposite connection is made the oxide coating quickly disintegrates and the condenser short-circuits the line to which it is connected.

In forming the aluminum oxide, current is forced through the condenser under an increasing voltage until this reaches about 425 volts. As the oxide film forms, the current through the condenser drops, and finally comes to a value of $\frac{1}{4}$ milliameter or less per microfarad of condenser. With the thickness of film so formed the capacity of the device is about 0.25 microfarad per square inch of electrode. Thus the standard 8 microfarad condenser has 32 square inches of aluminum used in its anode.

A condenser so formed will stand an operating voltage not to exceed 400 volts, and the voltage must never be allowed to reverse. This means that if a c.c. voltage of 300 is impressed on the condenser a ripple of not to exceed 100 volts (max) can be tolerated. Or, if 200 volts of c.c. voltage is used a ripple of not to exceed 200 volts (max.) can be tolerated. Both of these cases impress on the condenser a maximum of 400 volts.

The above values are for the electrolyte at about 25° C.; if this temperature rises, the maximum safe voltage diminishes, and the leakage current increases. After a condenser has been in service a long time the film improves to such an extent that the manufacturer guarantees a leakage current of only 0.1 milliampere per microfarad of capacity.

After standing some time with no polarizing voltage impressed the film suffers disintegration to some extent, so that when connected to a c.c. line the leakage current is large for a few minutes, as the film is repairing itself.

There is much loss in these condensers, apparently due to the electrolyte resistance; the power factor for the a.c. ripple superimposed on the c.c. polarizing voltage is sometimes as high as 70 per cent, indicating an equivalent series resistance as great as the capacitive reactance.

Occasionally laboratory work requires condensers for use on circuits of a thousand volts or more; neither waxed paper, nor electrolytic condensers, are suitable for this work. Mica, glass, oil, or air dielectric is used in such circuits. When in use on a.c. circuits, especially if the frequency is high, the solid dielectrics, even mica, warm up, and this decreases their puncture voltage. A mica condenser which operates satisfactorily on a 2000-volt c.c. circuit, for example, may break down in a very short time if 1000 volts of radio frequency is impressed.

If condensers are connected in series on a continuous, or pulsating, voltage line, a suitable leak resistance must be shunted across each one of the series, or they will probably all puncture. Thus, suppose two paper condensers, each good for 500 volts, are connected in series on a c.c. line. The voltage will divide according to their *insulation resistances*, and as they may differ 10 to 1 the voltage divides unevenly. If one has 1000 megohms and the other 100 megohms resistance, practically all of the 1000 volts of the line will be impressed across the good condenser, the one with 1000 megohms resistance. This excessive voltage will puncture this condenser, and then the total 1000 volts will appear across the poor condenser and, of course, it will break down. If, however, a resistance of say one megohm had been connected across each of the condensers, then each one would have had 500 volts impressed, and both would stand up to the voltage.

In case one condenser will not stand as much voltage as the other then a proportionately lower resistance must be connected across the poorer condenser, to divide the total voltage properly.

Variable Condensers.—Variable condensers generally use air or oil for dielectrics; one set of plates is fixed and another rotates into

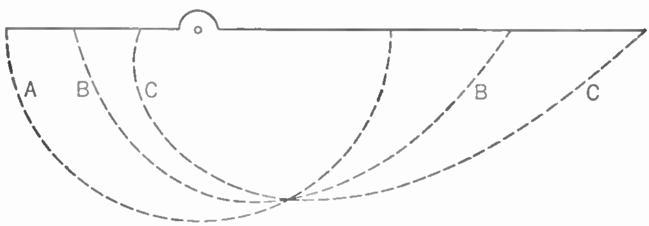


FIG. 19.

more or less interleaved relation with the stationary plate. If the rotatable plates are semicircular in form the capacity of the condenser varies linearly with the angle through which the plates are turned. This is the type of condenser used for ordinary laboratory work, and for standard condensers.

The rotatable plates may, however, be of such a spiral form that the relation between capacity and angle of rotation has any desired law.

Thus, we have logarithmic condensers, straight-line wave length, and straight-line frequency condensers. The SLW condenser is useful for wave meters and the SLF condenser is useful in the tuning condensers or broadcast receivers. In Fig. 19 are shown the approximate forms of the movable plates of the SLC, SLW, and SLF condensers. The exact form of the two latter depends upon the residual capacity of the circuit, that is, the total capacity in the tuned circuit when the movable plates are turned to their zero position.

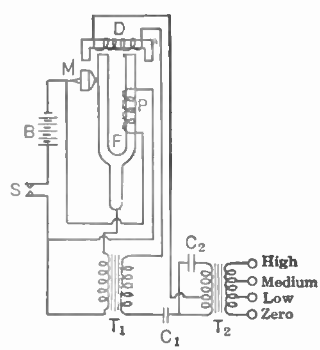


FIG. 20.

Power Supply.—The radio laboratory should have a variable frequency alternator, good for frequencies from 30 to 150 cycles per second, capable of generating 100 volts at the lowest frequency. Its wave form should be as pure as possible; less than 1 per cent departure from a pure sine wave is a reasonable specification. Its current capacity need not be greater than 5 amperes. This machine is suitable for the first nine experiments of the course outlined in this text.

For small amounts of power for bridge measurements a tuning-fork hummer is useful. In this scheme, the circuit of which is given in Fig. 20, a microphone button M is subject to the vibrational motion of fork F. A 6-volt battery B supplies the power to magnetize the fork by coil P, and to excite the primary of transformer T_1, through the button M. Fluctuations in this primary current set up alternating currents in the secondary circuit of transformer T_1, which alternating currents, acting through coil D, serve to maintain the fork in oscillation at its natural frequency.

The output transformer T_2, excited through condenser C_2, serves

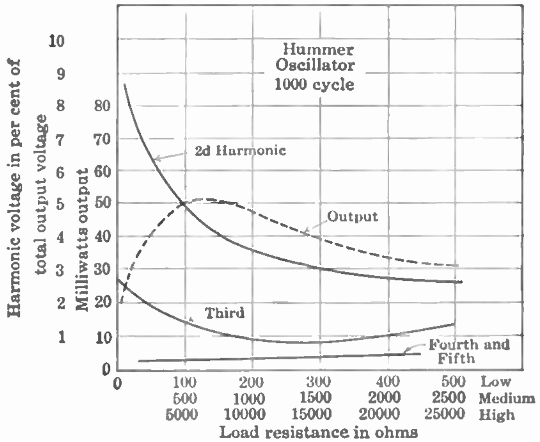


FIG. 21:

to purify the generated wave somewhat, making the output power from its secondary terminals reasonably sinusoidal in form. Fig. 21 shows how the output power varies in amount as the load circuit resistance varies. For each of the output terminals (low, medium, and high) different load resistances are required. Thus on the " low " tap with a load circuit resistance of 120 ohms, a power output of 50 milliwatts is available, the input being about 0.8 watt.

The terminals " zero-high " give about 20 volts on open circuit; these terminals will thus give plenty of voltage for operating the input circuit of a small power tube, to get watts of power, if such is desired.

The percentage of harmonic voltages present in the output is shown in Fig. 21. The principal harmonic is the second; with optimum load resistance, this is 5 per cent of the fundamental; the third harmonic is about 1 per cent, and the higher harmonics are negligible.

The appearance of this type of a.c. power generator is shown in

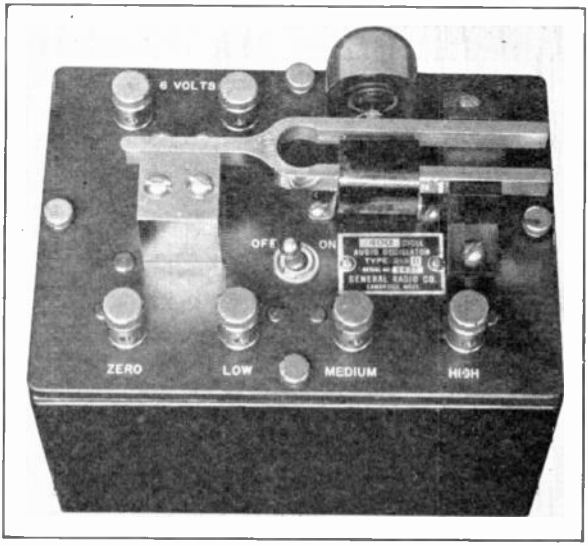


FIG. 22.

Fig. 22; it can be obtained for frequencies of 400 or 1000 cycles per second for about $25. Special ones can be built for frequencies from 100 to 1500 cycles per second.

For getting power over wider frequency ranges a vacuum tube

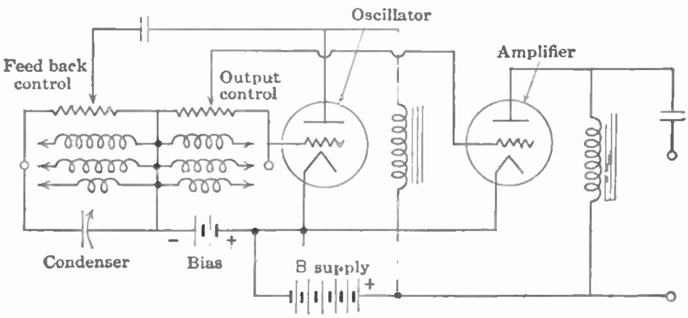


FIG. 23.

oscillating circuit is necessary; such oscillators are available which will give about one watt of power over a frequency range from 100 to 100,000 cycles per second. The circuit diagram of one such oscillator is shown in Fig. 23, and the general appearance is shown in Fig. 24. It is designed to feed into a load circuit of about 5000 ohms, and will impress on such a circuit a voltage about 25 per cent of the voltage used in the plate circuit. Such vacuum tube oscillators must generate harmonics, but by proper design these are kept down to about the same proportion they have in the hummer oscillator shown in Fig. 21.

To cover the wide frequency range, three different coils are used, and a condenser is required which can be set at any value from $1.0\mu f$

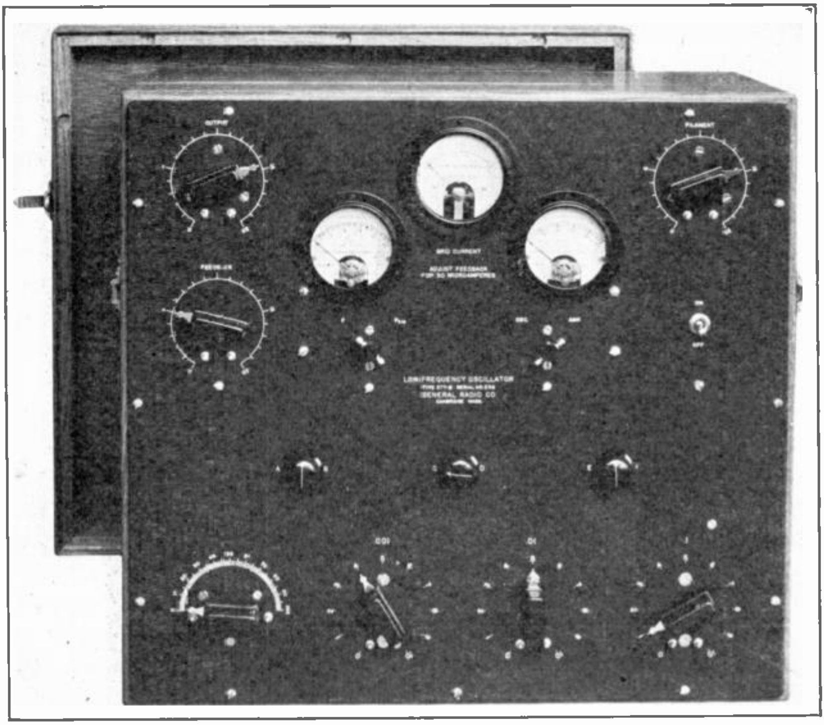


FIG. 24.

down to practically zero. Three decade condensers (0.1, 0.01, and $0.001\mu f$ steps) and an air condenser permit this variation.

The circuit used in this oscillator is given in Fig. 23, and the exterior and interior views are shown in Figs. 24 and 25.

Such an oscillator as that shown in Fig. 24 is necessary in the radio laboratory, in fact two or more are advisable. They require batteries for filament and plate circuits; such an oscillator costs about $350.

For radio frequency power one of the standard oscillating circuits may be used, such as are studied in Exps. 27 and 32. Several sets of coils are required to cover a wide frequency range; with a variable condenser of .003μf capacity, in parallel with about four .002μf fixed, mica condensers, (each fitted with a single pole switch to be put in or out of the circuit) one set of coils will cover a frequency range of about six to one.

It is well to mount the oscillator in a copper-lined box, to prevent

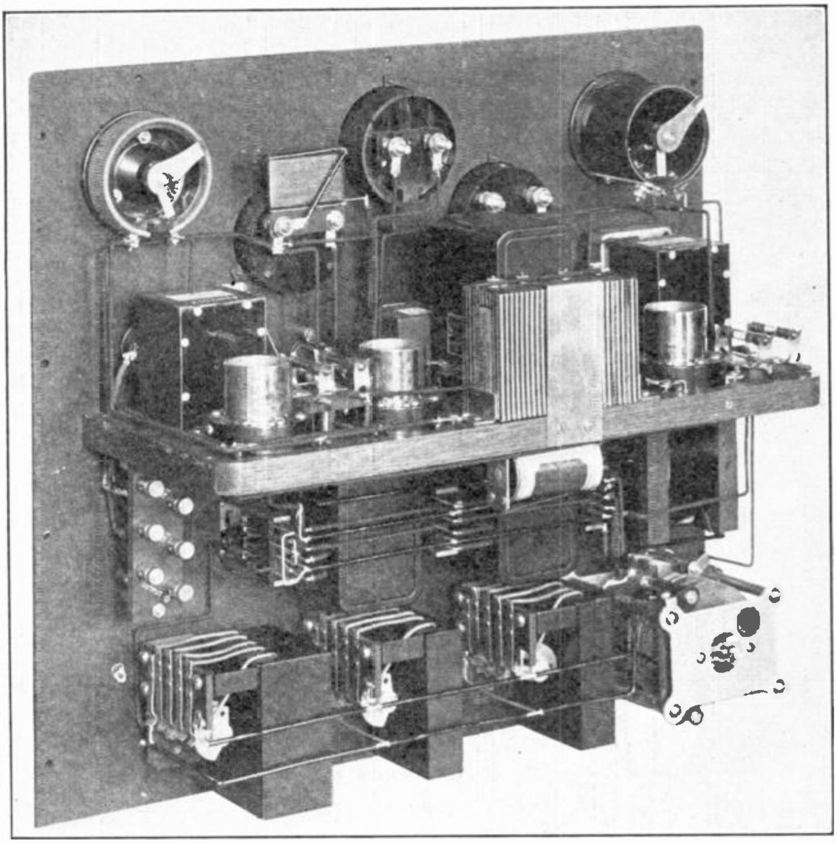


FIG. 25.

excessive interference with other radio frequency set-ups in the laboratory. The coil which supplies power to the load is preferably coupled to the oscillating circuit instead of being part of the oscillating circuit. An electrostatic shield between the oscillating circuit and this pick-up coil should be used, to prevent any but magnetically induced voltages being supplied to the load.

The pick-up coil should be adjustably coupled to the oscillating circuit, and this adjustment should be such as can be smoothly varied, when the copper-lined cover of the box containing the oscillator is closed. Suitable holes in the cover permit the operator to read meters in the grid and plate circuits, oscillating circuit, etc.

Another type of audio frequency oscillator uses the "beat frequency" principle. Two vacuum tube oscillators, of adjustable frequency difference, feed a detector and this supplies the input circuit of a power amplifier tube. Such a circuit scheme, arranged to get all of its power from the 60-cycle 110-volt power line of the laboratory has just appeared on the market (General Radio Co.).

EXPERIMENT 1

Object.—Measurement of coefficient of inductance and resistance of coils at low frequencies. Air core and iron core coils.

Analysis.—The various measurements are to be made by suitable ammeter, voltmeter, wattmeter, and frequency meter.

First, using an air core coil of a few tenths of a henry inductance, hold the frequency constant at a suitable value, and vary the current through a suitably wide range, reading volts, amperes, and watts.

The effective resistance is obtained from wattmeter and ammeter readings, the impedance from voltmeter and ammeter, and reactance is calculated from the known values of impedance and resistance.

In determining the effective resistance of the coil by this method, care must be exercised that the various meters are properly connected (with respect to each other), and that the power lost in the meters themselves is properly allowed for. A proper connection of the meters is shown in Fig. 26.

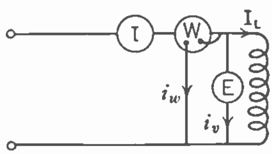


FIG. 26.

Suppose the coil tested has 5 ohms resistance and 0.2 henry inductance. At 80 cycles and 100 volts, one ampere will flow and there will be a loss of 5 watts in the coil. But the voltmeter itself will use about 8 watts, the potential circuit of the wattmeter perhaps 5 watts, and the ammeter and current coil of the wattmeter may use 2 more. So the meters themselves use 15 watts and the coil being measured uses only 5. Evidently proper connection of meters and allowance for their power consumption must be made if this method of measuring effective resistance is to yield results of any value.

In addition to the errors due to power consumption by meters, the actual reading of the wattmeter itself cannot be done with any degree of precision, unless a special meter is used. The wattmeter used in this test would probably have a current capacity of 1.5 amperes, a voltage rating of 150 volts, and a full scale reading of 150 watts. So the reading of 5 watts (power used by the coil) would be only one or two divisions.

For circuits of this nature, it is advisable to use special wattmeters, suitably calibrated for low power factor circuits. For example, by using a suitably weak spring the wattmeter mentioned above may have a full scale reading of say 20 watts. The 5-watt consumption of the coil would then make the meter read well up on the scale.

On account of the reactance of the potential circuit of the wattmeter, the current in this circuit lags behind the impressed voltage and so causes an error in the wattmeter readings.* This error is negligible with circuits of reasonably high power factors, but may be appreciable in low power factor circuits and must be taken into account. It results in a power indication greater than the actual power used in the coil.

Using the iron core coil (the 110-volt coil of a small transformer is suitable), carry out the same tests as were used for the air core coil, i.e., frequency constant and various current values, and current constant with various frequencies.

In the first of these runs it is advisable to start with very low currents, the maximum current used in the run being perhaps twice the normal exciting current of the coil. Thus a 3 K.V.A. transformer has an exciting current of about 2 amperes. The run then should consist of readings at about eight values of current, from 4 amperes down to 0.1 of an ampere or less. Results such as shown in Fig. 27 will be obtained.

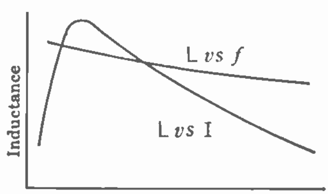


Fig. 27.

Providing a sufficient range of current is used, the iron core will show a low inductance (with the smallest value of current) which rises, with current increase, to a maximum and then decreases again. Unless very small currents are used the test will show inductance continually decreasing as the current is increased.

The air core coil should show inductance and resistance practically independent of current and frequency; the iron core coil should show variation, with current changes, as indicated in the above paragraph, and for fixed current and varying frequency, the inductance will decrease and resistance increase slightly as frequency is raised.

For tests of the kind called for in this experiment about eight properly spaced points should be obtained.

Plot the curves on cross-section paper.

* See page 39 (Fig. 31) for analysis of this error.

EXPERIMENT 2

Object.—Measurement of mutual inductance of air core coils, for various spacing and positions.

Analysis.—The two coils, the mutual inductance of which is desired, are connected in series, and the self-induction of the pair (treated as one coil) is determined as in Exp. 1. With the relative position of the two coils left unchanged, the connection of one of them is reversed so that the current through it is now opposite to what it was before, for a given direction of current in the first coil. The self-induction of the pair (again treated as one coil) is again determined.

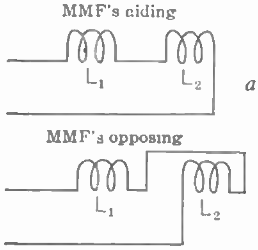


FIG. 28.

The mutual induction is equal to one-fourth of the two self-inductions, as determined above.

With the coils connected as in (*a*) of Fig. 28, the equation of reaction is:

$$e_1 = (R_1 + R_2)i + (L_1 + L_2 + 2M)\frac{di}{dt} \quad . \quad . \quad . \quad . \quad (5)$$

and with one coil relatively reversed as in (*b*) of Fig. 28, the equation of reaction is:

$$e_2 = (R_1 + R_2)i + (L_1 + L_2 - 2M)\frac{di}{dt} \quad . \quad . \quad . \quad . \quad (6)$$

If we designate $(L_1 + L_2 + 2M)$ by L' and $(L_1 + L_2 - 2M)$ by L'' and solve the above equation we find

$$M = \frac{L' - L''}{4} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

In another method for finding M, current of known magnitude and frequency is passed through the first coil and the voltage generated in the second coil is measured. Then M is calculated from the relation

$$E_2 = 2\pi f M I_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

In the formula E_2 is the voltage *generated* in the second coil; it may well be that the voltage *measured,* with the ordinary type of voltmeter, is appreciably less than the generated voltage. The ordinary type of iron vane voltmeter draws about 0.15 ampere when indicating full scale, and this amount of current, flowing through the impedance of the second coil, produces a quite appreciable internal drop. The voltage measured may possibly be as much as 5 per cent lower than the generated voltage. In case an electrostatic voltmeter of suitable range is available, this error can be avoided; this type of voltmeter draws so little current that no appreciable internal drop occurs.

In actually carrying out this test it is advisable to connect one coil

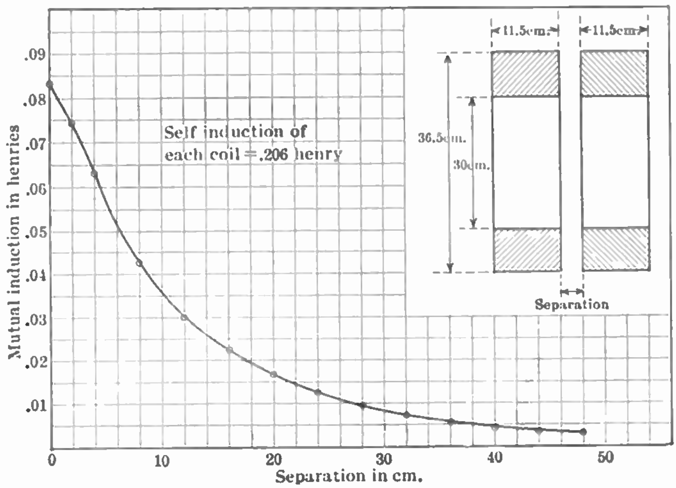


Fig. 29.

to the blades of a D P D T switch connected as a reversing switch. The various measurements can then be quickly obtained.

With two coils arranged to move apart while staying coaxial, measure sufficient values of M to construct a curve showing how M varies with separation. It will be found advisable to use small increments of separation when the coils are close together as M varies most rapidly here. Obtain values for increasing separation, until this is at least twice the diameter of the coils.

With the coils separated a distance somewhat less than the diameter of the coils, find how M varies with the angular relation of the two, by turning one of them on its axis. Take readings for every 10° rotation.

With the two coils separated sufficiently to permit the insertion of a third coil between the two, find the effect, on M, of the two coils being tested, of short-circuiting the intermediate coil; then close the intermediate coil through several values of resistance (up to a value equal to two or three times the reactance of the intermediate coil) and again measure the M of the pair of coils.

In general, it is the coefficient of coupling of two coils, rather than their mutual induction, which is desired. The value of this coefficient is obtained from the relation

$$K = \frac{M}{\sqrt{L_1 L_2}}. \quad \quad \quad (9)$$

Its maximum theoretically possible value is unity; actually in practice with single-layer solenoids fitting snugly one inside the other, the

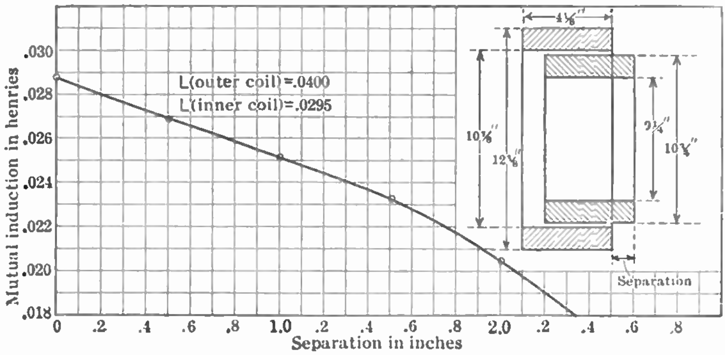


FIG. 30.

coefficient may be as high as 90 per cent. As the coils are moved out one from the other the coupling drops slowly at first (while one is still partly inside the other) and then rapidly as one is completely withdrawn.

In Figs. 29 and 30 are shown experimentally determined values for coils such as may be used in this test.

Plot values of M and K for the various runs, using M and K for ordinates, and separation, rotation, or intermediate coil resistance as abscissae.

EXPERIMENT 3

Object.—Measurement of insulation resistance, capacity, and power factor of condensers at low frequencies.

Analysis.—The ordinary condenser has such a small equivalent series resistance that practically no error is made in substituting its reactance for its impedance. This approximation yields the relation

$$I = 2\pi fCE \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

in which I is the charging current of the condenser having capacity C, when voltage E of frequency f is impressed upon it.

The equivalent series resistance of the condenser is obtained from the readings of the voltmeter, wattmeter, and ammeter, as is the effective resistance of a coil.

Using an alternator of the best wave form available, measure the capacity and series resistance of a bank of paraffin paper condensers, and find out how these vary with frequency (voltage held constant) and how they vary with voltage at a fixed value of frequency. Compare the measured value of capacity with that marked on the condensers.

In this condenser test it is extremely difficult to measure the equivalent series resistance because it is generally very low and the instrument corrections may be larger than the power the condensers are using. Thus let us consider a bank of 40 microfarads with 100 volts impressed, at such a frequency that $2\pi f = 500$. The charging current should be measured as 2 amperes, and the condenser reactance is 50 ohms. If now the power factor of the condensers is 0.5 per cent (General Radio Co. gives power factor of their mica condensers as 0.001–0.002 and of their paraffin paper ones as 0.01), the equivalent series resistance will be only 0.25 ohm.

Let us assume that the resistance of the potential circuit of the wattmeter is 1000 ohms and the inductance of this potential circuit is 4 millihenrys. The reactance of the potential circuit is then 2 ohms, and the current in the potential circuit lags behind the voltage impressed on the circuit by 0.114°.

The current and voltage relation in the circuit is then as shown in Fig. 31. The current into the condenser leads the voltage by (90°–

.285°), because, with the 0.5 per cent power factor assumed above, this phase relation is required.

The wattmeter should read $EI \sin 0.285°$, and it actually reads $EI \sin 0.171°$, which means that the power indicated by the wattmeter is only about 60 per cent of what it should be. Thus the equivalent series resistance of the condenser will be measured as only 60 per cent of its true value. As better condensers are used this error becomes increasingly important; in fact, with ordinary high-quality paraffin condensers, the wattmeter may read zero!

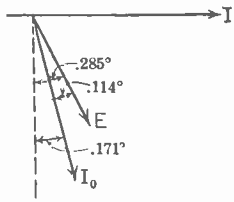


FIG. 31.

In addition to this error, those due to the power consumption of the meters (as discussed in Exp. 1) are also present.

It is likely that another error, of different type, may be encountered in the experiment. No alternator generates a really pure sine wave; the wave form has harmonic voltages superimposed, giving what are called the ripples, or teeth, on the voltage wave. When these ripples are present they always indicate that there are two harmonics present, in addition to the fundamental voltage. In a salient pole alternator the order of these harmonics is fixed by the number of armature teeth per pair of field poles. Thus an alternator having 12 teeth on the armature per pair of poles would generate, in addition to its fundamental voltage, the eleventh and thirteenth harmonics. Let us suppose such a voltage wave is used in this condenser experiment. Suppose the wave is given by the equation

$$e = 100 \sin \omega t + 9.1 \sin 11\,\omega t + 7.7 \sin 13\,\omega t. \quad . \quad . \quad (11)$$

Such a voltage wave, operating in the ordinary voltmeter, would give a reading of 71.1 volts, whereas if no harmonics were present it would read 70.7 volts; that is, the harmonics scarcely affect the reading of the voltmeter.

What current would flow on the condenser, when this voltage acted upon it? Each voltage would produce its own current, and this would have the same magnitude as it would have if the voltage were acting alone. Suppose the condenser has such capacity that the fundamental voltage alone causes a current of one ampere to flow. Then each of the harmonics will also cause one ampere of current to flow, and the condenser current will be made up of one ampere of fundamental current, one ampere of eleventh harmonic and one ampere of thirteenth harmonic. Such a current, flowing through the ordinary iron vane ammeter, would give a reading of 1.73 amperes!

If then we use the voltmeter reading and ammeter reading, in combination with the fundamental frequency, in the equation

$$I = 2\pi fCE$$

and solve for C, the value we obtain will be practically 73 per cent too high!

The amount of error caused by this effect depends upon how much the voltage wave is distorted; many laboratory machines in use today have waves more distorted than the one we assumed, and so this wave form error is present in most tests of this kind.

To eliminate this error it is only necessary to put a suitable choke coil in series with the alternator supply; the high frequency currents which cause this error can not readily flow through the choke coil and so are eliminated.

Great care must be observed, however, if the value of inductance of the choke, combined with the true value of the capacity being measured, is such as to satisfy the condition that

$$f = \frac{1}{2\pi\sqrt{LC}}. \qquad (12)$$

for any value of frequency to be used in the test. At the frequency where this relation is satisfied (resonant frequency), the voltage across the condenser may be ten or twenty times as great as the voltage the alternator is supplying, and unless care is exercised to keep the alternator voltage very low in the speed range where this frequency is generated, the condenser will be punctured by the excessive voltage set up in the circuit. The amount of voltage rise across the condenser (in excess of the alternator voltage) at this frequency depends upon the resistance of the choke coil used; the higher this is the less is the chance of breaking down the condenser.

Care must also be observed that the value of inductance used for a choke is not such as to satisfy the resonance relation for any harmonic of the alternator voltage, for any frequency to be used in the test. An excessive error will be experienced if such a condition is obtained; thus it is possible for the voltmeter and ammeter readings to indicate a value of capacity perhaps five times as great as it really is.

Very peculiar results may be obtained if the choke coil has a closed iron magnetic circuit, so it is best not to use such a coil unless necessary. If the magnetic circuit, however, has an appreciable air gap an iron core choke is satisfactory and even preferable to an air core coil.

The insulation resistance of the condenser may be measured by means of a megger, or if this is not available, an ordinary c.c. voltmeter. In general the resistance of such a voltmeter is given on the card which the manufacturer sends with the meter; in other cases, the resistance of the meter is given on the scale card of the meter as " 100 ohms per volt," or similar figure. Suppose a 150-volt meter having 80 ohms per volt is being used; its resistance is 12,000 ohms. Now the voltage of some suitable c.c. source (of proper voltage) is read directly by the voltmeter and is again read when the condenser is connected, in series with the voltmeter, to the same voltage source. The reading of the voltmeter for the two conditions, in combination with the known voltmeter resistance, gives the insulation resistance of the condenser.

If there were no leakage through the condenser the reading of the voltmeter would be zero, with the condenser connected in series with it; the fact that there is some reading shows that some current is leaking through the paraffin paper or other insulation.

If the reading of the voltmeter is V_1, when connected directly, and V_2 when connected in series with the condenser, and its resistance is R_v, then the insulation resistance of the condenser R_c is given by the relation

$$R_c = R_v \frac{V_1 - V_2}{V_2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (13)$$

Using paraffin paper condensers, make two runs to measure capacity and power loss in the condenser. In one run hold the frequency fixed (at approximately the middle value of the range available) and vary the impressed voltage on the condenser through as wide a range as feasible. Make another run holding the voltage fixed at some suitable value and vary the frequency through as wide a range as feasible. Get about eight sets of readings for each run.

In measuring the characteristics of the electrolytic condenser, special circuit arrangement must be made, as it is generally necessary to maintain a suitable value of c.c. polarizing voltage when making the a.c. measurements. In Fig. 32 the electrolytic condenser is shown at C. A milliammeter A_1, in series with a suitable choke coil L, in series with the variable source of voltage E, forms the polarizing circuit. For E small dry cells may be used, as the current taken is only a few milliamperes at most.

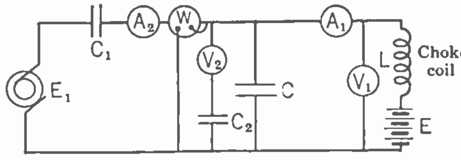


FIG. 32.

The alternating current circuit comprises the alternator E_1, in series with suitable meters and the paper condenser C_1. This condenser

C_1 serves to prevent the alternator from short-circuiting the source of polarizing voltage, E. In series with the alternating current voltmeter V_2 is a condenser C_2; this is to prevent the reading of this meter from being affected by the c.c. polarizing voltage in the condenser. The amount of this polarizing voltage is read by voltmeter V_1. Note that this is connected so that its current does not flow through the milliammeter A_1.

The wattmeter potential circuit is affected by the polarizing voltage, as well as by the voltage of alternator E_1. Its reading will, however, be unaffected by this condition because there is no continuous current flowing through its current coil. The wattmeter should be of the ordinary type, not the low power factor type called for in the previous

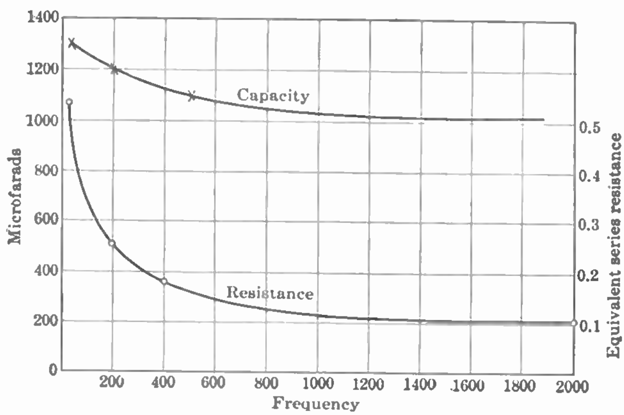


FIG. 33.

tests. The capacity of condenser C_1 (paraffin or oiled paper), as well as that of C_2, should be as large as feasible. The latter should be at least large enough to make its reactance (for the lowest frequency used in the test) not more than 10 per cent as great as the resistance of the voltmeter in series with it. If C_1 is too small the alternator will not be able to impress sufficiently high voltage on C, the condenser being measured; its reactance should be only a fraction that of C, for ordinary conditions.

The voltage of E, the polarizing agent, should always be at least 40 per cent greater than the effective a.c. voltage being impressed on the condenser. The reactance of the choke coil L should be large compared to the reactance of the condenser C, otherwise the ammeter A_2 will read too small a current. Its reactance should be at least fifty times as large as that of C.

Make one run to see how the capacity of C varies with voltage, frequency being constant. In this run set E at a value about 50 per cent greater than the highest value of alternating voltage to be used in the run.

With fixed suitable values of E_1 and E, make a run to see if the capacity and power loss of the electrolytic condenser varies with frequency, varying this throughout as wide a range as feasible.

With fixed frequency and fixed E_1 vary E to see if capacity, power loss, and condenser leakage current vary with polarizing voltage. Do not take readings, after changing E, until the milliammeter A indicates that the leakage current has reached a reasonably steady value.

In Fig. 33 are shown some characteristics of typical electrolytic condensers. Others available on the market show a much lower power factor than that given by these results.

Get about eight sets of readings for each of the three runs outlined above. Plot curves of capacity, power loss, power factor, and leakage current as ordinates, using for abscissae the factor which was varied in the run.

EXPERIMENT 4

Object.—Measurement of inductance and resistance of a coil, to which another coil, of various resistances, is magnetically coupled.

Analysis.—When a closed secondary circuit is coupled to a coil in which an alternating current is flowing, current will flow in this secondary circuit, the magnitude of which depends upon the induced voltage, and the reactance and resistance of the secondary circuit.

Of course, when current flows in the secondary circuit, power must be used here, and this power must come from the coil which, by the action of mutual induction, is setting up the current in the secondary circuit. If then we read the power supplied to the first coil, by a wattmeter, the reading of this instrument must give not only the power actually expended in the coil to which it is connected, but also the power being supplied to the secondary circuit. If then we use the customary definition of a.c. resistance,

$$R = \frac{\text{watts}}{I^2} \quad \ldots \ldots \ldots \quad (14)$$

in which the watts are given by the wattmeter reading, and I is given by the ammeter reading in the primary circuit, evidently R must be of complex nature, not only representing the actual resistance of the primary circuit itself but having an additional component which indicates *an increase in the effective resistance of the primary circuit,* due to the power supplied to the secondary.

The voltage induced in the secondary circuit is given by ωMI_1, and if the impedance and resistance of the circuit are indicated by Z_2 and R_2, respectively, the power supplied to the secondary circuit is

$$\left(\frac{\omega MI_1}{Z_2}\right)^2 R_2.$$

Hence the total power supplied to the primary circuit, and which of course is read by the wattmeter, is

$$I_1{}^2R_1 + \left(\frac{\omega MI_1}{Z_2}\right)^2 R_2$$

and if we divide the wattmeter reading by the square of the primary current we obtain the effective, primary resistance. It is indicated by R'_1 and evidently is equal to

$$R'_1 = R_1 + \left(\frac{\omega M}{Z_2}\right)^2 R_2. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (15)$$

The second part of the right-hand term must evidently be the increase in effective primary resistance due to the presence of the secondary circuit. It increases with the square of the magnetic coupling, decreases with increase in secondary impedance, and increases with secondary resistance (for a fixed secondary impedance).

The secondary current must have an effect upon the magnetic field set up by the primary coil, because of the mutual induction between the two coils. By considering the phase relations, it will be found that the secondary coil tends to *demagnetize* the primary coil, resulting in a decrease in the apparent reactance of this coil. The effective reactance of the primary coil is given by the expression

$$X'_1 = X_1 - \left(\frac{\omega M}{Z_2}\right)^2 X_2. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (16)$$

The second part of the right-hand term can never be greater than the term X_1 so that the reactance of the primary coil is always positive, and smaller than its true value, namely, X_1.

Evidently if the secondary circuit has no resistance, no power can be used in it with any finite value of current, so for very low secondary resistances the increase in primary resistance is small. And for very large secondary resistances the current flowing in the secondary circuit approaches zero, so the power (which varies with the *square of the current*) approaches zero. Hence for this condition also the secondary circuit will have a negligible effect on the resistance of the primary circuit. Some intermediate value of secondary resistance must therefore result in a maximum increase in primary resistance. By calculus it may be found that this critical value of secondary resistance should be obtained when the secondary resistance is equal to the secondary reactance, i.e., $R_2 = \omega L_2$.

Using the two coils, the mutual induction of which was measured in Exp. 2, arrange voltmeter, ammeter, and wattmeter to measure the power and power factor of one of them, which is to be connected to a variable frequency alternator, as indicated in Fig. 34. To the other coil connect a variable resistance which can be continuously varied from zero to a value about three times the value of the reactance of the second coil. Have a suitable ammeter in this circuit for measuring secondary current; use one the full scale reading of which is equal to the secondary current, when, no resistance is added in this circuit and tightest coupling obtains.

Do not change to a lower range meter for adjustments giving small value of current unless necessary; the ammeter resistance may be a quite appreciable part of the resistance of the secondary circuit, and, as the resistance of an ammeter varies inversely with the square of its current range (approximately), changing from one ammeter to another will materially change the resistance of the secondary circuit. Of course, if the ammeter resistance is known and allowed for, the ammeters may be changed at will.

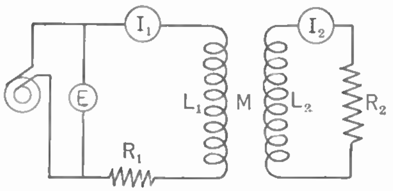


Fig. 34.

With tightest possible coupling and no added secondary resistance, adjust the alternator frequency to about its mid-range value and then increase the alternator voltage until the ammeters read well up on their scales; the wattmeter will read a small fraction of its scale, unless a special low power factor wattmeter is used.

Holding impressed voltage and frequency constant, vary the secondary resistance from zero to its maximum value, reading meters at about eight points. Repeat this run for medium coupling and for weak coupling.

With medium value of secondary resistance vary M in about eight steps (from maximum to a small value), keeping frequency and impressed voltage constant.

With tight coupling, and such value of secondary resistance as gives a secondary resistance equal to secondary reactance at the middle frequency, using a suitable value of voltage vary the frequency over the available range, keeping the voltage impressed on the primary coil constant.

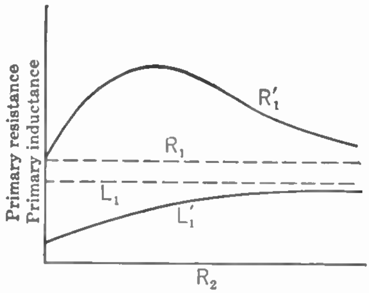


Fig. 35.

Calculate effective reactance and resistance of the primary circuit for all runs, as well as the reactance and resistance of the primary coil by itself. If these results are not available (from Exp. 1), make another run through the same frequency range as called for in the last run of this test with secondary circuit open; from these readings the true reactance and resistance of the primary coil by itself can be calculated.

In Fig. 35 are shown typical results obtained from runs performed as above.

Plot curves for all runs, using secondary resistance, mutual inductance, or frequency, for abscissae, according to which was varied in the run being plotted. Plot effective primary resistance and reactance as ordinates.

EXPERIMENT 5

Object.—To study the effect of a secondary circuit upon the primary circuit to which it is magnetically coupled, the secondary circuit having inductance, capacity, and resistance in series.

Analysis.—For this case, as for the one discussed in Exp. 4, the effective resistance of the primary circuit must be increased by the presence of the secondary circuit, and the amount of this increase must depend upon the amount of power used in the secondary circuit. The circuit arrangement is shown in Fig. 36.

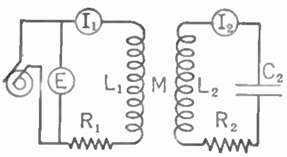


FIG. 36.

For the case considered in Exp. 4, the effective reactance of the primary coil was diminished by the presence of the secondary circuit, and this diminution occurred for all conditions that might obtain. In the case considered in this experiment, however, the reactance of the primary circuit may be increased, reduced to zero, or even made negative, according to coupling, frequency, etc.

The same pair of equations as given in Exp. 4 hold good for this case, also, but their analysis leads to different results than were previously obtained. We have

$$R'_1 = R_1 + \left(\frac{\omega M}{Z^2}\right)^2 R_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (17)$$

and for the present case we have

$$Z_2^2 = \left(\omega L_2 - \frac{1}{\omega C_2}\right)^2 + R_2^2. \quad . \quad . \quad . \quad . \quad (18)$$

It is evident that Z_2 may reduce to R_2 for this case; whenever L_2, C_2, and ω are related to make $\omega L_2 - \frac{1}{\omega C_2} = 0$, this will be true. When the frequency impressed on the primary circuit is that frequency which produces resonance in the secondary circuit (considered by itself) the expression for effective primary resistance is

$$R'_1 = R_1 + \frac{\overline{\omega M}^2}{R_2}. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

Evidently now the increase in effective resistance of the primary may be very great. As R_2 is diminished, the term $\frac{(\omega M)^2}{R_2}$ increases, resulting in large increases in R'_1.

This is the case frequently encountered in actual radio circuits; the effect of the secondary circuit on the value of R'_1 is so great that R_1 itself may generally be neglected! That is, practically all of the effective resistance of the primary circuit is due to the power transferred to the secondary circuit. Of the total power $I_1^2R'_1$ supplied to the primary circuit, the amount $I_1^2R_1$ is used to develop heat in the primary coil itself, and the amount $I_1^2\frac{\overline{\omega M}^2}{R_2}$ is transferred to the secondary circuit.

The effective reactance of the primary coil is given by

$$X'_1 = X_1 - \left(\frac{\omega M}{Z_2}\right)^2 X_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (20)$$

and as X_2 is zero at the resonant frequency, it is evident that when the frequency impressed on the primary is the resonant frequency of the secondary circuit (considered by itself) the effective reactance of the primary coil is the same as if the secondary circuit were absent.

Furthermore, it is evident that the effective reactance of the primary circuit (which is itself evidently inductive, that is, positive reactance) may even be negative. The formula for effective primary reactance may be written

$$X'_1 = X_1 - \left(\frac{\omega M}{Z_2}\right)^2\left(\omega L_2 - \frac{1}{\omega C_2}\right) \quad . \quad . \quad . \quad . \quad (21)$$

from which it appears that at frequencies lower than the resonant frequency of the secondary circuit the secondary reactance is negative $\left(\frac{1}{\omega C_2}\text{ is greater than }\omega L_2\right)$, and because of the negative sign before the second term the positive reactance X_1 is *increased* by the presence of the secondary circuit. This makes the primary circuit act as if the coil had a greater inductance than its true value. Because of the presence of Z_2 in the denominator of the term $\left(\frac{\omega M}{Z_2}\right)^2$ and because near the resonant frequency of the secondary circuit Z_2 becomes small, the effect of the secondary circuit on the primary reactance is greatest near the secondary resonant frequency.

Thus, at frequencies much below the secondary resonant frequency the effect of the secondary circuit on the primary inductive reactance is small, but what effect there is acts to augment the apparent inductance of the primary coil. As resonant frequency is approached, this augmentation is greatly increased, making the primary coil act as though it had five or ten times as much as its actual inductance. The magnitude of this effect evidently depends upon the coupling of the two coils, being greater as the coupling becomes tighter.

When the impressed frequency is the secondary resonant frequency the reactance of the primary coil is its true inherent reactance, the secondary circuit having no effect, as discussed earlier.

At frequencies slightly above the resonant value the term $\left(\frac{\omega M}{Z_2}\right)^2$ is still large and the sign of the term $\left(\omega L_2 - \frac{1}{\omega C_2}\right)$ has reversed, that is ωL_2 is greater than $\frac{1}{\omega C_2}$ so the parenthesis is positive. This then may result in X'_1 being less than X_1, and in fact, in most laboratory tests where low resistance coils and tight coupling are used, X'_1 is actually negative.

This means that the primary coil is acting as though it were a condenser; it actually draws a leading current from the generator to which it is attached. In Fig. 37 the analysis carried out above is depicted graphically. The curves marked R_1 and L_1 show the actual resistance and inductance of the primary coil *per se;* these are the values measured when the secondary circuit is opened. The curves marked R'_1 and L'_1 show the measured values of primary resistance and inductance when the secondary circuit is closed and so reacting on the primary. The difference between the two sets of curves, for any frequency, gives the value of resistance and inductance introduced into the primary coil by the reaction of the secondary circuit. Both of these increase as the coupling of the two coils is increased and as the resistance of the secondary circuit is decreased.

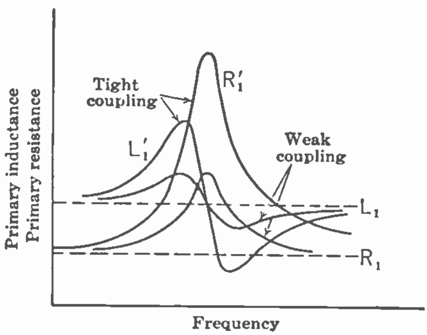


FIG. 37.

In carrying out the tests called for below, care must be exercised that the condensers in the secondary circuit are not punctured by excessive voltage set up in this circuit. When the impressed frequency is the resonant value for the secondary circuit, comparatively large currents may flow in this circuit, setting up voltages across the condenser much greater than the voltage being generated by the alternator. To protect against this condition we would naturally put a voltmeter across the condenser and take its reading. As will be discussed more in detail in the next experiment, this is not a safe procedure, and the condenser may be punctured even though the voltmeter never shows a voltage great enough to bring this about.

To guard against excessive voltage the reading of the ammeter in the secondary circuit must be depended upon. If the value of capacity has been properly chosen for the test, the secondary circuit will be resonant at the middle of the available frequency range. If this frequency and the value of the capacity being used, and the maximum safe voltage (effective a.c.) for the condensers, are known, the ammeter serves as a safeguard by noting that its reading must not exceed $2\pi f_0 CE'$, where f_0 is the resonant frequency, C is the capacity and E' is the maximum effective voltage the condenser will stand.

Thus suppose the secondary coil has an inductance of 0.2 henry and the middle value of frequencies available is 80 cycles. The proper condenser to use in the secondary circuit is obtained from the formula

$$80 = \frac{1}{2\pi\sqrt{0.2C}} \text{ or } C = 20\mu f.$$

Now if the maximum safe voltage is 200, the current in the condenser, given by the formula $I = 2\pi 80 \times 20 \times 10^{-6} \times 200$, must not exceed about 2 amperes. Hence in adjusting the alternator voltage the frequency should be set at the resonant value and the voltage then raised until the secondary ammeter reads about 2 amperes. Experimentally the procedure is to put a very low voltage on the primary coil, vary the frequency until the secondary ammeter shows its peak value, hold the frequency at this value and raise the alternator voltage until the secondary ammeter reads the desired value.

With the highest value of M obtainable, no added resistance in the secondary circuit, and the highest value of impressed voltage permissible (obtained as outlined above), vary the frequency throughout the available range, keeping the impressed voltage constant. Read primary volts, amperes, and watts and secondary current.

Repeat this run with medium coupling and again with loose coupling. Get about eight sets of readings for each curve.

Repeat the tight coupling run with a non-inductive resistance added to the secondary circuit; the value of this added resistance should be two or three times as much as that already in the secondary circuit.

With tight coupling, frequency held at its resonant value, impressed voltage held constant at its highest permissible value, read primary volts, amperes, and watts and secondary current as the resistance in the secondary circuit is varied from its minimum value (that inherent in the coil and ammeter) to a value about three times as great as the reactance of the secondary coil. Get readings for about eight different values of resistance.

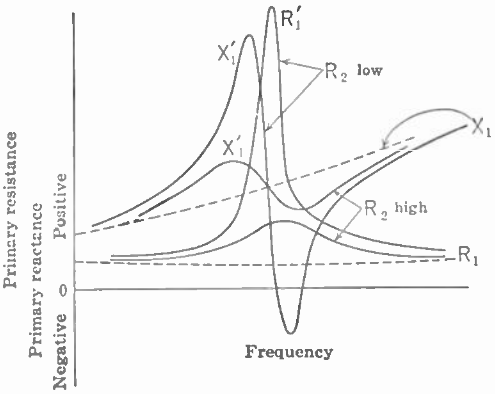


FIG. 38.

Plot curves for all runs, using for abscissae in each case the quantity which was varied. For the last run plot also the power supplied to the secondary circuit, using here also the values of secondary circuit resistance as abscissae.

In Fig. 38 are shown typical results for primary effective resistance and reactance, for a run such as called for above.

EXPERIMENT 6

Object.—To study resonance characteristics of a circuit containing inductance, capacity, and resistance in series. Effect of series and shunt resistance on sharpness of resonance. Determination of decrement of the circuit. Use of air core inductance and of iron core inductance. Effect of air gap in the iron core.

Analysis.—The coil and condenser are first properly selected so as to satisfy the resonance condition $\left(\omega L = \frac{1}{\omega C}\right)$ for a frequency at about the middle of the available frequency range. In general the coil will not be of variable inductance, so that the condenser will be the factor which is selected of proper magnitude to satisfy the resonance relation. Thus suppose an inductance of 0.5 henry of 10 ohms resistance is available, and that the frequency range available is from 50 to 100 cycles. The circuit should give resonance at about 75 cycles, and by calculation $C = \frac{1}{(2\pi f)^2 L}$ we find that a condenser of 10 microfarads is suitable.

The coil and condenser, in series with suitable voltmeter, wattmeter, and ammeter, are connected to the variable frequency alternator, as shown in Fig. 39. It requires some calculation to determine what range these meters should have.

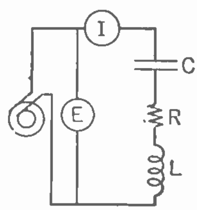


Fig. 39.

Let us suppose that the maximum safe working voltage (effective a.c.) for the condensers is 400 volts. We know that maximum current will flow in the circuit when resonance occurs, so we calculate how much current will flow in 10 microfarads of capacity, at 75 cycles, when 400 volts is impressed.

$$I = 2\pi f C E = 1.89 \text{ amperes.}$$

Hence the current coil of the wattmeter and the ammeter should be of 2 amperes capacity.

The reactance of the condenser at resonant frequency is 214 ohms, and if we assume the power factor of the condenser is $\frac{1}{2}$ per cent (a reasonable figure for good condensers) the equivalent series resistance of the condenser is about 1 ohm. The coil itself has 10 ohms (according to our previous assumption), so that the total series resistance of the circuit (excluding the meters) is 11 ohms.

Now at resonance there is no net reactance in the circuit, so that the impressed voltage has to overcome only the RI drop in the circuit. To force 1.89 amperes through 11 ohms resistance requires about 20 volts, so this is the value for which the alternator is to be adjusted.

And it must be borne in mind that even though the voltage impressed on the circuit is only 20 volts, at the resonant frequency there will be 400 volts across the condenser, and very close to 400 volts across the coil!

The values we have assumed for inductance, capacity, and resistance give about as sharp a resonance curve as can reasonably be expected at power circuit frequencies; in the average laboratory test the voltage across the condenser, at resonance, will not be 20 times as great as the impressed voltage, but more probably ten times as much.

In the previous test it was stated that an ordinary a.c. voltmeter was not suitable for protecting the condensers in such a test as this, and the statement will now be analyzed. Let us consider the circuit which has been treated so far in this test, namely, a circuit of 0.5 henry and 10 ohms resistance in series with 10 microfarads of capacity with an equivalent series resistance of 1 ohm, the impressed voltage being 20 volts and the impressed frequency the resonance value, namely, approximately 75 cycles per second. Previous calculation showed a voltage across the condenser of 400 and a current of 1.89 amperes at resonant frequency.

Now what will happen if a 400-volt iron vane voltmeter is connected across the condenser, with the idea of measuring this 400 volts? The resistance of the voltmeter will be about 6000 ohms, so that if 400 volts is impressed on it 0.066 ampere will flow through it. This amount of current, at a pressure of 400 volts, represents a power loss in the voltmeter of about 25 watts. This relatively high power consumption is quite common in voltmeters of the iron vane type.

Now such a power loss in a closed box the size of an ordinary voltmeter would result in excessive temperature in the box, if allowed for appreciable time. The temperature rise would give a large error in reading and quite possibly would injure the insulation in the meter winding. Because of this condition, voltmeters of this type are always equipped with a button switch so that the meter circuit is closed only when it is being read. So the question we have to solve is this: if the voltage across the condenser is 400 volts when the voltmeter circuit is closed, what is it when the voltmeter circuit is opened?

To answer this question we consider the voltmeter as a resistance across the condenser (which of course it is), and to make the calculation of the problem simple, we first change this shunt resistance into an equivalent series resistance. The equivalent series resistance is one which, connected in series with the condenser, will experience the same power loss as does the actual shunt resistance.

The current flowing into the condenser is given very nearly by the expression $I = 2\pi fCE$, and this current flowing through a series resistance R_s would produce a power loss of I^2R_s. Now the actual shunt resistance R_{sh} experiences a power loss of E^2/R_{sh}. So to get the equivalent series resistance we put

$$\frac{E^2}{R_{sh}} = I^2R_s \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (22)$$

and substituting the value of I previously given,

$$(2\pi fCE)^2R_s = \frac{E^2}{R_{sh}}$$

or

$$R_s = \frac{1}{(\omega C)^2R_{sh}} = \frac{X_c^2}{R_{sh}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (23)$$

Now the reactance of the condenser in the circuit being analyzed was previously found to be 214 ohms, so we have

$$R_s = (214)^2/6000 = 7.8 \text{ ohms.}$$

Hence putting the voltmeter across the condenser has the same effect on the circuit as would result from inserting 7.8 ohms resistance in series with the condenser.

Now such a resistance, in series with the 11 ohms already in the circuit, gives a circuit resistance of 18.8 ohms. With such a resistance, the alternator voltage, to give 400 volts across the condenser at resonant frequency, would have to be $18.8 \times 1.89 = 35.5$ volts.

When the voltmeter switch is opened, the circuit has only 11 ohms resistance. Such a resistance, with 35.5 volts impressed, would give a current in the circuit of 3.24 amperes, and such a current would give a voltage across the condenser of 690 volts!

Therefore, in this test if, with the voltmeter connected across the condenser, the alternator is adjusted (at resonant frequency) to give 400 volts across the condenser, there will be 690 volts across the condenser as soon as the voltmeter circuit is opened.

Conditions similar to this are quite common in radio circuits, and the student must be constantly on the lookout for them; connecting a meter to measure a certain quantity may of itself alter that quantity greatly so that the value measured is far from the value existing before the meter is connected.

Returning now to the general consideration of the problem, we note that in this circuit we have

$$I = \frac{E}{\sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}} = \frac{E}{\sqrt{R^2 + X^2}}. \quad . \quad . \quad . \quad (24)$$

At frequencies below resonant frequency, $\left(\frac{1}{\omega C}\right)$ is greater than ωL, the capacitive reactance predominates, and whatever current flows will lead the impressed voltage. At resonant frequency there is no reactance, so the current is in phase with the voltage. At frequencies higher than the resonant value the inductive reactance predominates and the current lags. At resonance the impedance is a minimum, so the current is a maximum (constant impressed voltage assumed). The values of current, plotted against frequency as abscissae, give the well-known resonance curve, shown in Fig. 40. The lower the resistance of the circuit, the sharper is this resonance curve. In Fig. 40 there are shown the curves of " volts across coil " and " volts across condenser," as well as the current in the circuit.

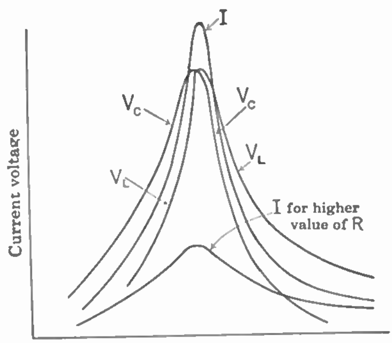


FIG. 40.

The radio engineer frequently uses the term " decrement " when discussing resonance phenomena. This quantity, generally represented by the letter small delta, δ, gives an accurate measure of the sharpness of the resonance curve and, furthermore, it tells the radio engineer how rapidly electrical oscillations will die away in the circuit if some disturbance sets up such a current and it is left to die out.

If f_r is the resonant frequency of a circuit consisting of L, C, and R in series, the decrement is defined by the ratio

$$\delta = \frac{R}{2f_rL}. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (25)$$

It is evidently a quantity somewhat akin to the power factor of the coil, at resonant frequency. We know that power factor $= \cos \phi = \frac{R}{Z}$, and if the resistance of a coil is small compared to its reactance we have $\cos\phi = \frac{R}{2\pi f L}$, so that we have the relation $\cos\phi = \frac{\delta}{\pi}$. This relation holds good only at the resonant frequency of the circuit. Now the quantity δ can also be obtained from the shape of the resonance curve. Referring to Fig. 41 at f_r (resonant frequency) the current is I_r. Now two other points are selected on the resonance curve where the current

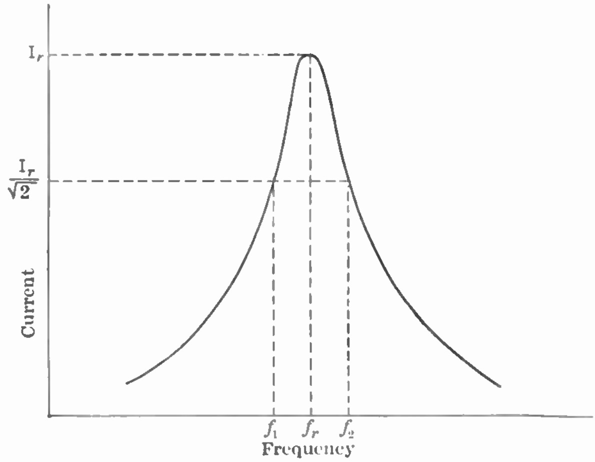


FIG. 41.

is equal to $I_r/\sqrt{2}$. These two currents occur at frequencies f_1 and f_2. It can be shown* then that

$$\delta = \frac{\pi (f_2 - f_1)}{f_r}. \qquad (26)$$

It is then possible to obtain the decrement from experimental results in two different ways—by measuring R, f_r, and L, and by measuring the frequencies f_r, f_1, and f_2 from the resonance curve.

Iron core inductances give most peculiar and unexpected results when used in such an experiment as this one. Entirely different results will be obtained if the frequency is started low and then raised to the higher value, than if it is started high and lowered gradually. The peculiar behavior is a result of the dependence of the iron permeability upon the current through the coil, and, of course, in this resonance run the current varies greatly.

Typical values for L and R are given in Fig. 42. The coil was the 95-volt winding of a 3KVA lighting transformer. In Fig. 43 are shown

* Principles of Radio Communication, 2nd ed., p. 77.

the resonance characteristics of a circuit made up of this coil and a 120-μf condenser. Starting with low frequency (25 volts held constant in the circuit), the current was so low that the inductance and resistance values of the iron core are on the left side of the maxima given in Fig. 42. As frequency increased the current increased, L increased (see Fig. 42), and the resonant condition was approached. When about 0.6 ampere was flowing (frequency 18 cycles) the maximum value of L occurred, if Fig. 42 is to be relied upon, but even so the inductive reactance was much less than the condenser reactance. Now as the frequency was raised and the current increased above 0.6 ampere, the *inductance began to decrease*, making the resonant frequency higher.

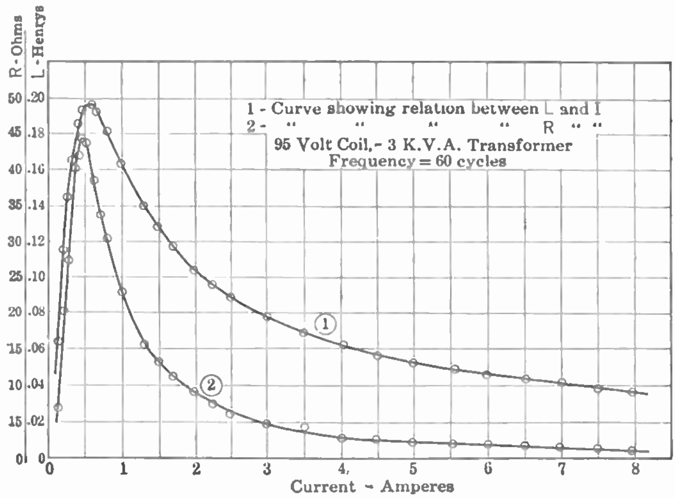


FIG. 42.

And as the frequency increased still further, L continually decreased (see Fig. 42), preventing the circuit from coming into resonance $\left(\omega L = \dfrac{1}{\omega C}\right)$. This continued until a frequency of 65 cycles was reached and the current rose to 8 amperes; for this condition the equality of inductive and capacitive reactances is nearly reached. Suddenly with the slightest increase in frequency the current drops from 8 amperes to 0.3 ampere, so that evidently the right-hand side of the resonance curve is unstable.

What happens is this. Just after the resonant condition is reached, the current, having passed its maximum value, starts to fall. But Fig. 42 shows that as the current decreases from a value of 8 amperes the inductance increases. But increase in inductance means a lower resonant frequency, and this action results in still lower current. So a higher inductance exists, resonant frequency for the circuit is still further reduced, and on the curve of Fig. 43 when the alternator was generating 70 cycles the resonant frequency of the circuit was about 40 cycles!

So with increasing frequency we see the effect of the varying permeability of the iron is to keep the resonant frequency of the circuit

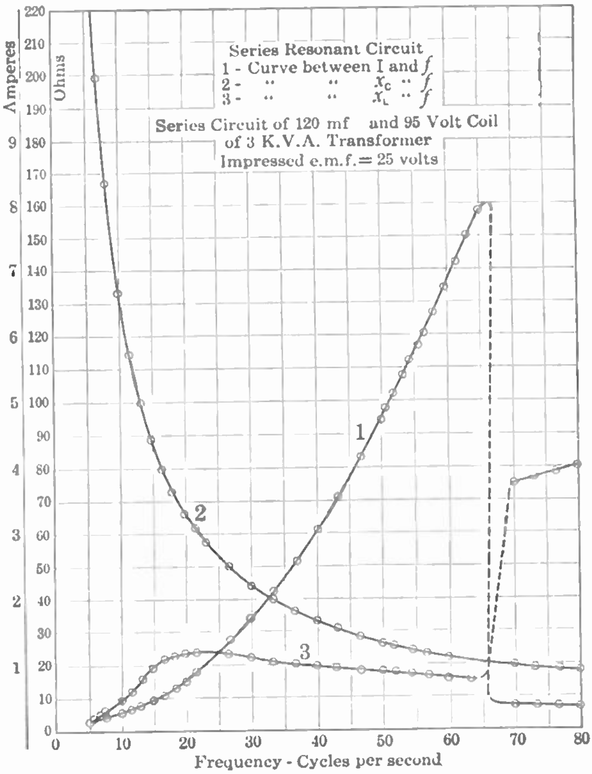


Fig. 43.

somewhat above the value of the impressed frequency; there comes a limit to the circuit's ability to maintain this condition and suddenly the resonant frequency of the circuit shifts backward and drops to a value below the frequency impressed on the circuit.

If an air gap is inserted in the iron core the instability of the circuit is correspondingly diminished, and with an air gap an appreciable fraction of an inch the unstable condition does not exist.

The curves of Fig. 42 illustrate an effect which the student finds difficult to appreciate at first; namely, an air core coil can be built with much lower power factor than an iron core coil. A transformer core is, of course, well laminated and has little eddy current loss, but even so the ratio of its reactance to resistance is only about three to one, whereas an air core coil of the same weight as this iron core coil (iron plus copper) would have a reactance fifteen or even twenty times as much as its resistance.

Perform runs as follows, getting sufficient points, properly spaced, to construct accurately the resonance curves.

With air core coil, no added resistance, with sufficient impressed voltage to give maximum safe voltage on the condenser at resonance, read current for various frequencies, holding voltage constant throughout the run.

Repeat with two values of added resistance, one about the resistance of the coil and the other about three times as much. For these runs increase the alternator voltage sufficiently to give the same maximum current for all three curves.

When these curves are plotted the current values are to be prorated to the values they would have had if the impressed voltage had been the same for all three runs. It is better to carry out the tests as suggested rather than actually to hold the voltage the same for the three runs, because in this case the currents would be so small as to be scarcely readable on the ammeter used for the first run. And it is generally advisable not to change meters in a set of runs like this because the ammeter itself has appreciable inductance and resistance (in the low range meters) and so the circuits for the several runs might have appreciably different values of inductance, when presumably they are the same.

With no added series resistance, and same impressed voltage as in the first run, make another run with a shunt resistance across the condenser using a resistance about ten times the value of the condenser reactance.

With a closed iron core inductance, and sufficient capacity to presumably give resonance at the mid-frequency range, try to get a resonance run showing the phenomenon illustrated in Fig. 43. (The inductance of a transformer core can be approximately calculated by assuming a reasonable exciting current for rated voltage and frequency and calculating the inductance from the approximate relation $2\pi fLI_{exc} = E_{rated}$.) If available, use an iron core coil with an air gap for getting another resonance run.

For all runs plot curves of current, inductive reactance drop, and capacitive reactance drop, against frequency as abscissae.

From the known impressed voltage, and maximum current from the resonance curve, calculate the resistance of the circuit ($R = E/I_{\text{resonance}}$). From this resistance and known value of inductance and resonant frequency calculate the decrement. Compare with the value of decrement calculated from the shape of the resonance curve.

For the run using a shunt resistance across the condenser, calculate the series resistance equivalent to the shunt resistance (at resonant frequency), and from this and the known resistance of the rest of the circuit get the total equivalent series resistance of the circuit. Check this value from the measured values of E and I at resonance. At resonant frequency calculate the ratio of reactance to resistance of the coil.

EXPERIMENT 7

Object.—To study parallel resonance and the effect of circuit changes on it. Effect of alternator wave form on accuracy of results. Effect of supplying power to a portion of the coil only.

Analysis.—In the previous resonance test it was necessary to observe caution to prevent dangerously high voltages being set up across the condenser. When the coil and condenser are connected in parallel, as called for in this test, the resonance characteristics are just as pronounced as they were for the series connection, but no dangerous condition is likely to be obtained when carrying out this test. The condenser never has a higher voltage than that impressed by the alternator.

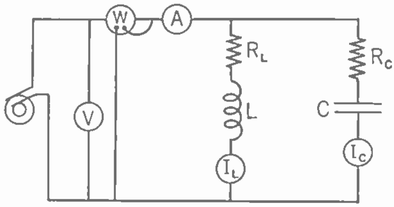


FIG. 44.

In addition to the ammeter, voltmeter, and wattmeter used in the line from the alternator to the parallel circuit, there should be used a suitable ammeter in each branch (Fig. 44).

When the power factor of both the coil and the condenser are low (as should always be the case in resonance tests) the resonant condition in the circuit occurs at practically the same frequency as it would if the coil and condenser were connected in series, namely, $f_0 = \frac{1}{2\pi\sqrt{LC}}$.

Resonance is defined for a parallel circuit in the same way as for a series circuit, namely, that frequency for which the circuit shows no reactance, in other words, that frequency for which the power factor of the circuit is unity, or we may say, that frequency for which the line current is in phase with the line voltage.

Let us suppose that the coil we are to use has 0.2 henry inductance and 5 ohms resistance, and that the mid-frequency available is 80 cycles. We have $2\pi fL = 125$, and the condenser required to give a reactance of 125 ohms at 80 cycles is 20 microfarads. Suppose we decide to use 125 volts as the line voltage and the frequency is varied from 40 to 120 cycles.

The first thing to note about this circuit is that *each branch takes* *exactly the same current and power as it would if the other branch were not present.* Such a statement could not be made, of course, for the series connection of the last test.

The current in the two branches will change with frequency (impressed voltage being held constant at 125), as shown in Fig. 45. The condenser current is a linear function of the frequency and the inductance current is nearly a hyperbola. The lower the coil resistance, the closer the curve approaches a rectangular hyperbola.

For the first analysis we will suppose that the condensers are so good as to have a negligible power loss; this means that the condenser current leads the line voltage by 90°.

If the coil is properly built of cable its resistance will be essentially constant at its c.c. value, in the frequency range used.

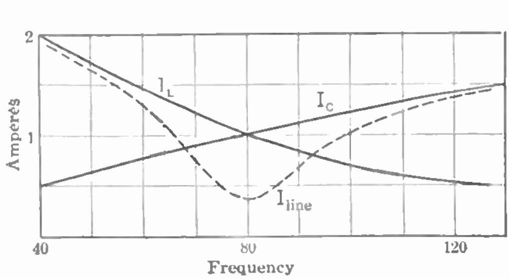


Fig. 45.

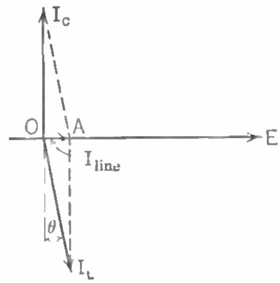


Fig. 46.

At the resonant frequency both branch ammeters will indicate 1 ampere, as nearly as can be read. (This neglects the important factor of alternator wave form which will be discussed later.) The reactance of the condenser is 125 ohms, the reactance of the coil is 125 ohms, and its resistance is 5 ohms. This small resistance, compared to the reactance, means that the coil impedance is sensibly the same as its reactance. This is the reason both branches draw the same current, even though one has 5 ohms resistance and the other has none.

The line current will be the vector sum of the two branch currents, and this is shown in Fig. 46. The two branch currents combine to give a line current OA, much smaller than either branch current. The current OA is evidently equal to the current I_L multiplied by $\sin \theta$ and this is sensibly the same as $I_L \tan \theta = I_L \frac{R_L}{X_L}$. Substituting the proper values of R_L and X_L gives a line current OA of .04 ampere. The line impedance is given by the ratio of line voltage to line current or $125/.04 = 3125$ ohms. Strange as it may seem, this is the fact, that this coil of only 5 ohms resistance, when shunted by a suitable condenser, acts like a resistance of thousands of ohms! And stranger still, it is evident upon further analysis that, as the resistance of the coil is diminished, the apparent line resistance becomes higher and higher. From inspection of Fig. 46 it is seen that as the coil resistance becomes smaller the current OA becomes less, even though it remains in phase with the line voltage. But when the line current becomes smaller for a fixed line voltage, we must conclude that the resistance of the circuit has become larger.

The power actually used in the coil is $I_L^2R_L$, and, of course, the line resistance must be such a quantity that it, multiplied by the square of the line current, gives the power used by the coil.

The line current is sensibly equal to I_LR_L/X_L, and this current, squared, multiplied by the line resistance, which we will call R', must be equal to $I_L^2R_L$. So

$$I_L^2 \frac{R_L^2}{X_L^2} R' = I_L^2 R_L \quad \text{or} \quad R' = \frac{X_L^2}{R_L} \quad . \quad . \quad . \quad . \quad (27)$$

This relation holds good, of course, only at the resonant frequency, when the circuit draws a current in phase with the line voltage. Also it assumes a coil and condenser, the impedance of which is sensibly the same as the reactance; this is always the case in actual radio circuits.

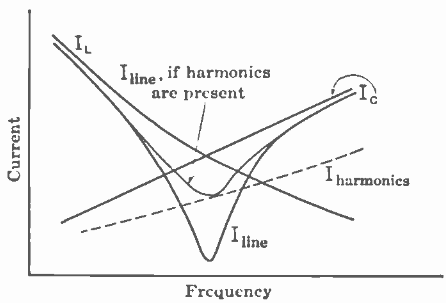


Fig. 47.

By constructing Fig. 46 for a various frequencies, and transposing the line current so obtained to Fig. 47, we see that the line current is in the form of a V, being practically the same as the larger branch current for frequencies distant from the resonant frequency, and having very small values (perhaps one-tenth of either branch current) in the region of resonant frequency.

Now it will be found, when this test is carried out in the laboratory, that instead of verifying the theory outlined above, the results disprove it completely. The circuit we have analyzed, which should have an effective resistance of 3125 ohms at resonance, will probably show a resistance of 30 ohms only, possibly less. This, it will be seen, is not an error measured in a few per cent, but in many hundred per cent.

The reason for this great discrepancy between the experimental results and those predicted by theory, lies in the wave form of the alternator. If this generates a pure sine wave, theory and experiment will agree, but, of course, no generator can produce a really pure sine wave. A voltage wave which has so little distortion as to appear truly sinusoidal will still give very poor agreement between theory and experiment. We will analyze a typical case.

Let us suppose an alternator having 12 teeth per pair of poles, a common construction. This machine, with salient poles, will generate an eleventh and a thirteenth harmonic of nearly equal amplitudes. We will suppose each of them is 5 per cent of the fundamental; this would produce a ripple on the voltage wave (or " teeth " on the voltage wave as they are frequently called) which would have at certain parts of the wave, an amplitude 10 per cent of the amplitude of the fundamental. At other parts of the voltage wave, the amplitude of the ripple would be practically zero. (This waxing and waning of the amplitude of the ripple is due to the thirteenth harmonic " catching up " with, and passing, the eleventh harmonic twice per fundamental cycle.) Such a wave does not look badly distorted, and is much better than is given by the average laboratory generator.

Now, let us again suppose the voltage of the fundamental wave is 125, as before, so that the harmonics will each have a value of 6.25 volts. The coil we have been considering has 0.2 henry and 5 ohms, and the condenser has 20 microfarads of capacity and negligible resistance.

The current, of fundamental frequency, is one ampere in both coil and condenser, and the line current of this frequency is 0.04 ampere, as shown on page 63. The line current has this very small value because the two reactive currents, that in the coil lagging 90° and that in the condenser leading 90°, just neutralize one another.

Now it is a fundamental proposition in electrical theory that if several voltages of different frequencies are impressed upon a circuit, each voltage will produce the same current as it would if it were acting alone. This is known as the principle of superposition. It holds good only if the L, R, and C of the circuit are constant, independent of voltage and current intensity. It thus does not apply to circuits having iron core coils, wires of varying temperatures, microphonic contacts, and the like. In our case the coils and condenser are constant in their inductance and capacity, so we may use this principle of superposition.

Now turning our attention first to the circuit branch containing the coil, we conclude that the upper frequencies in the voltage wave will produce current of negligible magnitude. Thus, the eleventh harmonic will produce a current of 0.0045 ampere, and the thirteenth harmonic a current of 0.0038 ampere. Compared to the 1.00 ampere of fundamental current in the coil, these currents are entirely negligible.

But when we consider the capacity branch of the circuit we find an entirely different state of affairs. The eleventh harmonic produces a current of 0.55 ampere and the thirteenth harmonic a current of 0.65 ampere. These currents are nearly as large as the fundamental current; their unexpected magnitude comes about because of the high frequency. This factor acts to offset the comparatively low magnitude of the voltages.

Now these two currents, 0.55 ampere of eleventh harmonic and 0.65 ampere of thirteenth harmonic, are not neutralized by corresponding currents in the coil, so they must be supplied by the line itself. Thus the line must supply a current of 0.04 ampere of 60-cycle current, 0.55 ampere of eleventh harmonic and 0.65 ampere of thirteenth harmonic.

These three currents flowing through the line ammeter will produce thereon a reading of $\sqrt{.04^2 + .55^2 + .65^2} = 0.87$ ampere. That is, the line ammeter reads 0.87 ampere, when the theory (based on sine wave voltage) predicts a current of 0.04 ampere!

Now the power used in the parallel circuit is the same as if the voltage were a pure sine wave, because the upper harmonic currents flow only in the condenser branch, where we have assumed a negligible series resistance. Hence, the power loss in the parallel circuit, which is read in the line wattmeter, will be 5 watts, as it was before. The line resistance will be

$$R' = \frac{\text{watts}}{I^2} = \frac{5}{.87^2} = 6.85 \text{ ohms}$$

Or if we assume that the resistance may be obtained as before, namely, by dividing the line voltage by the line current, we find

$$R' = \frac{125}{.87} = 144 \text{ ohms}$$

So that this circuit, even at resonant frequency, does not give the same value of impedance when calculated by the two methods. The wattmeter method gives 6.85 ohms, the impedance calculation gives 144 ohms, and theory indicates that the resistance should be 3125 ohms.

The harmonic currents not only upset the resistance calculations but also spoil the sharpness of the resonance curve. In Fig. 47 are shown the fundamental currents through the coil and condenser I_L and I_C and the line current I_{line} due to these two currents. It shows a sharp minimum of only 0.04 ampere at the resonant frequency. In this figure is shown also the curve I_{har} which gives the harmonic current flowing in the condenser path. In the dotted line is shown the line current the experiment yields; instead of dropping to 0.04 ampere at the resonant frequency it falls only to 0.88 ampere.

To eliminate this error in performing the test on parallel resonance, it is necessary to place in series with the alternator a choke coil, and in making connections to the circuit under test the choke coil is treated as part of the alternator circuit. It is well to use an iron core choke with a small air gap in the magnetic circuit, if such is available. This choke coil then tends to eliminate the high frequency currents not only by its reactance but also by resistance due to core losses; these are much larger for the harmonics than they are for the fundamental frequency.

Referring to Fig. 48, which shows the connection to be used in this test, it can be seen that other resonances besides that being investigated may occur. For the high frequency harmonics being discussed the main circuit of L and C in parallel will be a capacity for all the frequencies to be used in this run. Thus, if the frequency variation in this test is to be from 40 to 120, the resonant frequency of the parallel circuit being, say, 75 cycles, then for the harmonics the parallel circuit will be a condenser, even at 40 cycles. In fact, for the eleventh harmonic the parallel circuit will be a capacity until the fundamental frequency goes lower than 7 cycles.

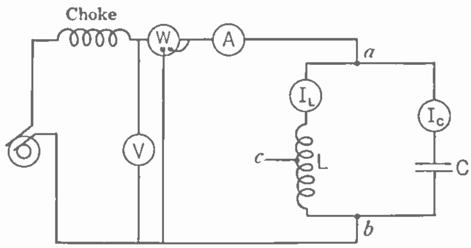


FIG. 48.

It is thus evident that the choke coil, in series with the parallel circuit (which can be replaced in our minds by a condenser), may show *series resonance* for one of the harmonic frequencies. This will bring an unexpected hump in the resonance curve. The series resonance will result in an exaggeration of the harmonic voltage across the condenser of the parallel circuit. Generally this phenomenon will not be encountered; if it is, a larger choke coil should be used and this will depress the series resonant frequency (for the harmonic) below the frequency range used in the test.

With the main inductance having a value of 0.2 henry, as has been assumed in this analysis, a choke coil of a few hundredths of a henry (say 0.05 henry) will be found satisfactory.

Besides investigating the phenomenon of parallel resonance in the circuit of Fig. 48, this test also involves a study of what happens if the power supply is made to only a portion of the coil, say from *a* to *c* of Fig. 48. The parallel circuit now has an inductance in one path and an inductance in series with a condenser in the other path. It will be found that the circuit shows resonance (line current in phase with line voltage) at just the same frequency as it did before, in fact this is a special case of the general theorem that when circuit reactances are large compared to resistances, *a network will show resonance at the same frequency no matter what two points of the network are used for supplying the power.* In the special case we are considering it will be found that the results are about as shown in Fig. 49. Curve 1 gives the line current when all of the coil is in one branch and only the condenser in the other. Curve 2 shows the line current when the power is supplied across half the coil only. It is supposed that the line voltage is the same for both runs.

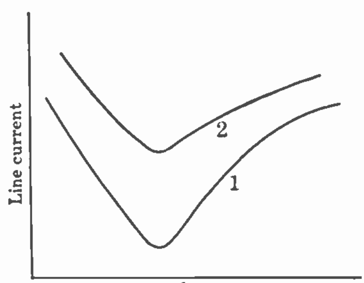


FIG. 49.

These curves are qualitative only; their exact shape, one compared to the other, depends upon how much mutual induction there is between the two parts of the coil. If the coupling between the two parts is weak, or if only a small part of the coil is used across the power line, large voltages may be built up across the condenser at some frequency used in the run; the reading of the ammeter in the branch containing the condenser will serve to indicate whether a dangerous condition exists, because the voltage across the condenser, the current in the branch, and the frequency being used, bear the relation mentioned several times before; i.e., $I = 2\pi fCE$. From preliminary calculations the student should know how much current can flow in the condensive branch of the circuit without danger of the condenser breaking down.

With the circuit arranged as in Fig. 44, make a run reading all meters, keeping impressed voltage constant. Add in the condenser branch a resistance about 10 per cent of the condenser reactance, and repeat the run.

With no resistance added to the original circuit, make a run for a small range of frequencies in the region close to resonance, after having put a suitable choke coil in series with the alternator. To hold the voltage on the test circuit constant during this run, it will be necessary to vary the alternator voltage considerably. The voltage in the circuit is equal to the voltage of the alternator minus (vectorially) the drop in the choke coil, and this will vary a great deal in magnitude and phase for the different frequencies used in the test.

With the power supply connected to one end of the coil and a tap about midway between the two ends, and no added resistance, make another run. Use the same impressed voltage as for the other runs, if the apparatus permits.

If time allows, get an oscillograph record of the branch and line currents, at the resonant frequency, for the various arrangements tested.

Plot curves and explain results obtained.

EXPERIMENT 8

Object.—Study of resonance phenomena in a circuit resembling that used in a radio frequency amplifier.

Analysis.—In general the circuits used in actual radio apparatus are somewhat more complicated in their arrangement and action than are those so far studied. The phenomena so far studied are involved in the action of the radio apparatus, but generally, not simply as studied in the foregoing tests, but in combination with other actions. The average tuned radio frequency amplifier involves circuits similar to those given in Figs. 50 and 51. The circuit of Fig. 50 is similar to that of the ordinary screen grid radio frequency amplifier, and that of

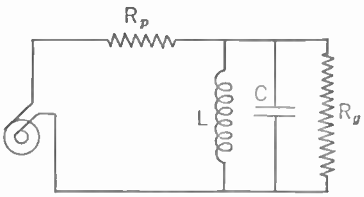


Fig. 50.

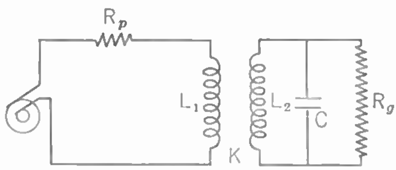


Fig. 51.

Fig. 51 represents one stage of the ordinary tuned radio frequency amplifier, used with the three-electrode tube. In each figure R_p represents the intèrnal a.c. resistance of the plate circuit of the vacuum tube, and R_g represents the a.c. resistance between the grid and cathode of the succeeding tube. The value of R_g is determined by grid bias, whether external grid leak is used, etc. The value of R_p depends for its value upon the type of tube, grid bias, and plate voltage used.

In this test we shall not take the values of inductance and capacity used in the actual radio circuits, because at the low frequencies given by an alternator resonance could not be obtained with microhenrys of inductance and milli-microfarads of capacity, such as the radio frequency amplifier uses. But we shall try to select such values of inductance, capacity, and resistance as will show, reasonably well, the performance of the actual radio apparatus.

In interpreting the result of this test, as well as the preceding and following experiments, it must be remembered that the resonance characteristics of actual radio circuits are much more pronounced than we can obtain at power frequencies. The sharpness of resonance depends upon the ratio of reactance to resistance of the coils and condensers used; whereas the coils of radio apparatus have a reactance one hundred times as much as their resistance, it is not feasible, at power frequencies, to get a reactance more than about 25 times the resistance. Even this ratio calls for a well-built coil weighing about 100 lbs.

With allowance made for this quantitative difference between the two cases, however, these low-frequency tests show very well the performance of actual radio apparatus.

In trying to set up a circuit, for the low-power frequency, which shall duplicate the performance of that of Fig. 50 used at radio frequencies, we note that in the radio circuit the value of R_p may be 200,000 ohms, the reactance of the coil, as of the condenser, may be 800 ohms, and R_g may be from 100,000 to 500,000 ohms. The power factor of the coil may be 1 per cent, so that from equation 23 on p. 55 we find that at resonance the equivalent resistance of the parallel circuit (neglecting for the moment the effect of R_g) is

$$R' = \frac{800^2}{8} = 80{,}000 \text{ ohms.}$$

Now if we use a coil with 0.2 henry and a resonant frequency of 80 cycles, the reactance of the coil will be about 100 ohms. If it has 5 ohms resistance, and a suitable condenser is used in parallel to establish resonance, the equivalent resistance of the parallel circuit is

$$R' = \frac{100^2}{5} = 2000 \text{ ohms.}$$

In the actual radio circuit the resistance R_p is about three times that of the parallel circuit, so for our test we should use $R_p = 3 \times 2000 = 6000$ ohms.

And for R_g we should use a resistance about the same as R_p, or say 5000 ohms.

To show the resonance phenomena we can measure either the current in the condenser, or the voltage across the parallel circuit. But if it is attempted to use an ordinary voltmeter for measuring the voltage it must be remembered that its resistance (for a 150-volt scale) will probably be only 2000 ohms, and this is in parallel with R_g which we have taken as 5000 ohms! If the characteristics of the circuit are to be maintained, therefore, it is necessary to use an electrostatic voltmeter for measuring the voltage across the condenser.

However, the resonance phenomena can be measured by the current in the condenser, and from this current, together with the known capacity and frequency, the voltage can be calculated.

The proper capacity to use in our test will have 100 ohms reactance at the resonant frequency (because we have assumed a coil which has this reactance), so that if 100 volts occurs across the parallel circuit at resonance, there will be one ampere of current flowing in the condenser. Thus an iron vane ammeter of 1-ampere range will be suitable for showing the occurrence of resonance.

Of course, it must be remembered that even putting a 1-ampere meter in series with the condenser has a really appreciable effect upon the power factor of the condenser path. Thus its resistance will be about 1 ohm, and this will increase the power factor from the value of possibly $\frac{1}{2}$ per cent (condenser itself) to $1\frac{1}{2}$ per cent. In fact, putting this 1-ampere meter in series with the coil has almost as much effect on the condenser power factor (hence upon the sharpness of resonance) as putting the resistance R_g in parallel with the condenser.

Putting 5000 ohms in parallel with a resistance of 100 ohms is the same as putting $X^2/R_{sh} = 100^2/5000$ ohms in series. This gives an equivalent series resistance of 2 ohms. Thus putting the ammeter in series is about the equivalent of decreasing the shunt resistance R_g from 5000 to 3000 ohms!

If available, it is well to use an electrostatic voltmeter across the condenser to indicate resonance; portable electrostatic meters with a full scale reading of 150 volts are readily obtainable.

To get a hundred or more volts across the parallel circuit will require a voltage on the alternator of about 400 volts. It will generally be necessary, therefore, to use a step-up transformer between the alternator and test circuit, as the average laboratory alternator will not give over 100 volts at the lower frequencies.

To measure the line current in this test it will be necessary to use a milliammeter because the current taken by the circuit is very small. Thus with a value of R_g of 6000 and equivalent resistance of the parallel circuit (at resonance) of 2000 ohms, the circuit has 8000 ohms resistance, and with 400 volts impressed the current will be 0.05 ampere.

It will, of course, require a rather sensitive ammeter to measure this small current, and the meter (if of the iron vane type) will probably have several hundred ohms resistance. Such a high resistance in an ammeter would greatly upset conditions in an ordinary circuit where ammeters are used, but in this circuit it is no detriment at all. The circuit requires a high resistance, and so the ammeter may be regarded as part of this resistance.

The circuit shown in Fig. 51 is that of the ordinary transformer repeating tuned radio frequency amplifier. Again R_p represents the resistance of the plate circuit of the tube, and R_g represents the resistance of the input circuit of the succeeding tube. In the actual radio circuit L_2 is about 250 microhenrys, C varies from 25 to 250 micro-microfarads, L_1 is about 10 microhenrys, the coefficient of coupling is about 75 per cent, R_p is about 5000 ohms, and R_g is about 100,000 ohms or more.

The resistance of L_1 itself is negligible, but the resistance which the secondary circuit introduces into the L_1 circuit may be quite large. From the analysis given in Exp. 5 we know that the value of resistance introduced into the primary circuit, at the frequency which gives resonance in the secondary, is $(\omega M)^2/R_2$.

From the data given above, $M = .75\sqrt{250 \times 10} = 37.5\mu h$, and for a frequency of 10^6 we have $\omega M = 236$ ohms.

A reasonable resistance for the coil L_2 at this frequency is 15 ohms, so that

$$\frac{(\omega M)^2}{R_2} = \frac{236^2}{15} = 3700 \text{ ohms.}$$

In other words, the resistance which the tuned secondary circuit introduces into the primary circuit is about the same magnitude as the resistance of the plate circuit of the triode. In the average radio set it is not as great as this, being perhaps one-quarter as great as the resistance of the triode plate circuit.

Let us suppose that in our test (Fig. 51) we select for the coil L_2 one having 0.6 henry inductance and 10 ohms resistance. The value of capacity required for tuning the secondary circuit to 80 cycles is about 7 microfarads. The reactance of the coil, as of the condenser, at resonant frequency is about 300 ohms. If we use for L_1 a coil of .03 henry, coupled 70 per cent to coil L_2, we shall have for the value of M,

$$.70\sqrt{.6 \times .03} = .094 \text{ henry}$$

and for ωM about 47.0 ohms. Then for $(\omega M)^2/R_2$ we have

$$\frac{47^2}{10} = 220 \text{ ohms.}$$

In order for the circuit of Fig. 51 to represent the condition in a radio circuit, therefore, we should use for the value of R_p about 500 ohms.

Some caution must be observed in the circuit or else the condenser may be punctured by a high voltage set up at the resonant frequency.

If we assume 300 volts as the maximum safe voltage (effective) on the condenser, the permissible current in the circuit at resonant frequency is obtained by the relation

$$I = 2\pi fCE = 2\pi 80 \times 7 \times 10^{-6} \times 300 \cong 1.0 \text{ ampere.}$$

With this current flowing the power loss in the circuit is 10 watts, and this is the amount of power the primary circuit must supply to the secondary. In other words, this is the maximum amount of power which may be supplied by the (square of the primary current) flowing through the resistance which the secondary circuit introduces into the primary. So we have I_p^2 (primary current) $\times$ 220 = 10 or

$$I_p^2 = .0455 \text{ or } I_p = .21 \text{ ampere.}$$

Now the total resistance of the primary circuit is 500 + 220 = 720 ohms, and to force a current of 0.21 ampere through this resistance requires about 150 volts. Hence this is a suitable voltage to use in running the test, for the conditions we have outlined.

Using the circuit shown in Fig. 50 with suitable value of voltage, held constant at the alternator, take readings of line current, condenser current (or condenser voltage) with no resistance shunted around the condenser, and two more runs with two values of shunt resistance, one representing a low resistance for the triode input circuit and the other representing a high resistance for the input circuit.

Perform similarly three runs on the circuit of Fig. 51, using values about as outlined in the analysis.

In both arrangements of circuit the curve between primary current and frequency is nearly a straight horizontal line, having more or less of a dip for that frequency which sets up resonance in the secondary circuit.

EXPERIMENT 9

Object.—To study resonance phenomena in two magnetically coupled circuits, both circuits having condensers in series with the coils.

Analysis.—The circuit we investigate in this experiment is shown in Fig. 52; in each circuit there are a coil and condenser in series, and the two circuits may react on one another through the mutual magnetic field, indicated by M.

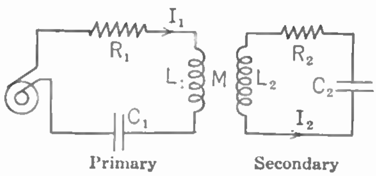


FIG. 52.

We will first consider the simplest possible case, i.e., $L_1 = L_2$ and $C_1 = C_2$, with R_1 and R_2 sufficiently small in comparison to L_1 and L_2 that they play no part in determining resonance conditions.

From Exp. 5 we know that the secondary circuit will affect both the resistance and reactance of the primary circuit, making the effective resistance and reactance of this circuit,

$$R'_1 = R_1 + \left(\frac{\omega M}{Z_2}\right)^2 R_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (28)$$

and

$$X'_1 = X_1 - \left(\frac{\omega M}{Z_2}\right)^2 X_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (29)$$

As R_2 is small compared to the reactance in the secondary circuit, it is apparent that the second circuit will introduce a large resistance into the primary circuit only when the term $\left(\frac{\omega M}{Z_2}\right)$ is large, and this occurs only for the condition that $Z_2 = R_2$. For this condition, as we have shown before, $R'_1 = R_1 + (\omega M)^2/R_2$ and that this holds good only when the frequency impressed on the primary is the resonant frequency for the secondary circuit. For frequencies appreciably different from this frequency the effective resistance of the primary circuit does not have values greatly in excess of its own true resistance. It may be twice as much, but it will not be ten or twenty times as much, as it may be for the resonant frequency.

In general it is the reactance which limits the current in such circuits as this, so we now investigate the effective reactance of the primary circuit, as given in eq. (29). First it is evident that, for the special circuit conditions we have specified, when $\omega L_1 = \dfrac{1}{\omega C_1}$ making $X_1 = 0$, we also have $\omega L_2 = \dfrac{1}{\omega C_2}$ or $X_2 = 0$. But if both X_1 and X_2 are zero the value of X'_1 (given in eq. (29)) is zero, so that the primary circuit must show the resonant condition, that is, line current in phase with line voltage.

But, even though the reactance is zero, it seems probable that the current in the primary circuit will not be large, because of the large resistance, introduced by the secondary circuit. For all values of coupling except the very weakest, this term will be very large compared to R_1 and so will result in low values of primary current (for a given impressed voltage) even though the circuit shows no reactance. And, of course, a low primary current results in a low induced voltage in the secondary circuit, and so a low secondary current.

Now evidently there may be other frequencies which show zero reactance in the primary circuit, because of the form of eq. (29). If we put $X'_1 = 0$ and simplify the solution of the resulting equation by neglecting the resistance terms in comparison to the reactance terms, we get a comparatively simple bi-quadratic equation which can be solved.

If we call $1/\sqrt{L_1C_1} = 1/\sqrt{L_2C_2} = 2\pi f_0$ and if we call $M/\sqrt{L_1L_2} = k$, then the equation yields for the two values of frequency for which the primary circuit shows zero reactance;

$$f' = \frac{f_0}{\sqrt{1 + k}} \qquad \qquad (30)$$

and

$$f'' = \frac{f_0}{\sqrt{1 - k}} \qquad \qquad (31)$$

These two resonant frequencies, f' and f'', not only show zero reactance, but they also show comparatively small resistance (when the circuit is analyzed more carefully) so that these two frequencies show minimum values of primary impedance. They, therefore, give the frequencies for which the primary circuit shows maxima of current, and are called the two resonant frequencies of the circuit.

It will be noticed that one of these frequencies is lower than the natural frequency of the circuit (either circuit, as the two have the same natural frequency) and the other occurs at a frequency higher than the natural frequency. Is there a physical explanation for these two frequencies which, in so far as this analysis has shown, are merely two roots of the quadratic equation obtained by putting the primary reactance equal to zero?

To find the reason for these two frequencies let us consider what effect the secondary circuit has upon the primary (which is connected to the variable frequency alternator as shown in Fig. 52) as the frequency is increased from low values.

For low frequency the secondary circuit shows high reactance, $\frac{1}{\omega C_2}$ being very large compared to ωL_2. A small current will, therefore, flow in this circuit, but what current does flow will *lead* the secondary voltage and hence will help to magnetize the magnetic circuit of L_1. (The student of electrical machinery will recognize this as the same principle which states that a leading current drawn from the armature of an alternating current generator helps the field winding to magnetize the field poles. And shortly we shall use the complementary principle that a lagging current in the armature circuit tends to demagnetize the field poles.)

Now if L_1 acquires more interlinkages per ampere (owing to the assisting action of L_2), then *it has a higher effective self-induction than it would have if L_2 were not present.* The amount of this increase depends upon the current in L_2 and upon its angle of lead.

As the frequency impressed on the primary is increased (hence, of course, that set up in the secondary also) the current in the primary circuit increases, and that in the secondary circuit increases, so that its effect on L_1 is increased. At some frequency (perhaps much below the natural frequency of the primary circuit acting alone) this augmented value of L_1 is sufficient to produce resonance in the primary circuit, and this is the frequency denoted by f' in eq. (30).

The amount by which this frequency f' differs from f_0, the natural frequency of the primary circuit, depends upon how much L_1 is augmented by the action of coil L_2, and this in turn depends upon how much current is set up in L_2. But this in turn depends upon how much voltage is set up in the secondary circuit, and this depends upon the coefficient of coupling between the two circuits. We can then see, at least qualitatively, the reason for the form of the equation $f' = f_0/\sqrt{1 + k}$ in which k is the coefficient of coupling of the two circuits.

Now at the frequency f' a maximum of current will flow in the primary circuit, and hence a maximum of induced voltage will result in the second circuit. As the constants of the second circuit have been assumed the same as those of the primary, a maximum current will flow in the secondary circuit at the same frequency at which the maximum induced voltage occurs. Hence, that frequency which shows a maximum of current in the primary, namely, f', will also result in a maximum of current in the secondary circuit. This statement is not quite true, because the resistance of the secondary circuit plays a minor role in determining the secondary maximum current, and the effect of the resistance is to make the secondary maximum current occur at a frequency slightly higher than that which gives a maximum primary current.

If now the frequency is still further increased, we soon reach that for which the secondary reactance is zero, and here the current of the secondary is in phase with its generated voltage and so neither magnetizing nor demagnetizing action is exerted on L_1 by L_2. Hence L_1 appears to have its true value. Now as we have assumed that $L_2C_2 = L_1C_1$ it is evident that when the secondary circuit shows resonance (current in phase with voltage) the primary circuit also shows resonance, that is, no net reactance.

There will not be a maximum of primary current at this frequency, however, because of the amount of resistance which the secondary circuit introduces into the primary, namely, $(\omega M)^2/R_2$. For low resistance secondary coil, and any value of M except the extremely low values, the value of this expression $(\omega M)^2/R_2$ is so great as to cause a *minimum of primary current* even though the reactance is zero.

But if the primary circuit has minimum current, the induced voltage in the secondary is a minimum and hence the secondary circuit also shows a minimum current. Here again the truth of this statement depends upon the coupling having some appreciable value, say for ordinary coils greater than perhaps 5 per cent.

Now as the frequency impressed on the primary is further increased, the current in the primary (and hence that in the secondary) begins to increase, because the secondary introduces less resistance into the primary. Thus the secondary coil begins to exert an appreciable magnetic effect upon L_1; but now the action is a demagnetizing one, because at frequencies above the natural one the current in the secondary circuit lags behind its generated voltage. Thus L_1 now appears to have a value of self-induction *less than its true value.*

As the frequency impressed is further increased, thus further diminishing the resistance introduced in the primary circuit by the secondary, the primary current increases, the secondary current increases, and the demagnetizing action on L_1 reaches still larger values.

Soon this action gives such a diminished value of L_1 that its reactance, combined with that of C_1, gives a net reactance in the primary circuit of zero and hence a maximum of primary current flows. And, just as before, when this occurs, a maximum current also occurs in the secondary circuit.

This simple analysis explains, in a rather superficial manner, to be sure, the form of resonance curves in a pair of coupled, tuned circuits. For three different values of coupling, there are given in Fig. 53 the forms of primary and secondary current for the arrangement of Fig. 52. The solid lines give the primary current and the dotted line the secondary current. Those marked 1 are for tightest coupling, and those marked 3 are for weakest coupling.

It will be noticed that primary and secondary maxima do not quite coincide, and that this departure from coincidence becomes greater

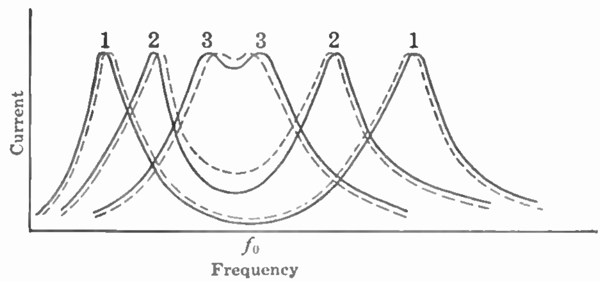


Fig. 53.

as the coupling is diminished. It will further be noticed that the two resonant frequencies are not equally spaced from the natural frequency of the two circuits. The higher frequency moves away from f_0 much faster than does the lower; as a matter of fact, for the tightest coupling imaginable (100 per cent) f' has moved from f_0 to $f_0/\sqrt{2}$ and f'' has moved from f_0 to infinity.

It will be noticed that outside the two resonant frequencies f' and f'' the primary current is larger than the secondary, and that in between f' and f'' the reverse is true. The latter part of this statement must be qualified and, in fact, reversed when the coupling becomes very weak, as we will now show.

At the frequency f_0 the resistance introduced into the primary circuit by the secondary is equal to $(\omega M)^2/R_2$. The resistance of the primary alone is R_1, so the total power supplied to the two circuits, by the alternator, is $I_1^2\left(R_1 + \frac{(\omega M)^2}{R_2}\right)$. The amount of power expended in the primary is $I_1^2R_1$, and the amount of power expended in the secondary is $I_1^2 \frac{\overline{\omega M}^2}{R_2}$. Now it is also true that the power expended in the secondary circuit is $I_2^2R_2$, and as we have assumed in the discussion that the two circuits were identical (i.e., $R_1 = R_2$), it follows that if $\frac{\overline{\omega M}^2}{R_2}$ is greater than R_2 then I_2 must be greater than I_1. Also it follows that if $\overline{\omega M}^2/R_2$ is equal to R_2 the two currents must be equal, and if $\overline{\omega M}^2/R_2$ is less than R_2 then I_2 must be less than I_1.

It thus follows that if $\omega M = R_2$ the current in the secondary (at frequency f_0) is just equal to the current I_1; if ωM is greater than R_2 then I_2 is greater than I_1, and vice versa.

From this discussion, it follows that for the tests shown in Fig. 53 at even the weakest coupling the mutual reactance was greater than the secondary resistance.

In case the circuits are mistuned, that is, L_1C_1 is not equal to L_2C_2, the above simple analysis does not suffice to predict the results. Mathematical analysis,* however, shows that in this case the two resonant frequencies are given by

$$f' = \sqrt{\frac{f_1^2 + f_2^2 + \sqrt{(f_1^2 - f_2^2)^2 + 4k^2f_1^2f_2^2}}{2(1 - k^2)}} \quad . \quad . \quad . \quad (32)$$

$$f'' = \sqrt{\frac{f_1^2 + f_2^2 - \sqrt{(f_1^2 - f_2^2)^2 + 4k^2f_1^2f_2^2}}{2(1 - k^2)}} \quad . \quad . \quad . \quad (33)$$

For this mistuned condition it also may be shown, by mathematical analysis, that the two maxima of current, at f' and f'', are not equal, as they are for the tuned condition. Unless the student is especially interested in the case it is sufficient to analyze the circuits experimentally, without attempting the mathematical analysis.

In Fig. 54 are shown the results of a test on mistuned circuits. In dashed curve (A) is shown the resonance curve of the primary for the tuned condition; the two resonance peaks have the same height. By increasing the capacity in the secondary about 100 per cent, both resonant frequencies are lowered, as common sense would predict, and the higher frequency f'' has a much greater current than the lower frequency f' (C). By increasing the capacity of the primary, instead of the secondary, the same values of f' and f'' are obtained (B), but now the greater current flows at the lower resonance frequency.

With tuned circuits identical (preferably with the same constants as used in Exp. 6) make three resonance runs, with maximum obtain-

* Principles of Radio Communication, 2d ed., pp. 119–120.

able coupling, minimum of about 5 per cent, and an intermediate value. Get sufficient points accurately to delineate the curves This calls for a closer spacing of the measured values in the regions of resonance than elsewhere.

For these runs the voltage impressed on the primary circuit may be twice the value which was found safe for the single circuit of Exp. 6. (This statement holds good only for couplings which make ωM greater

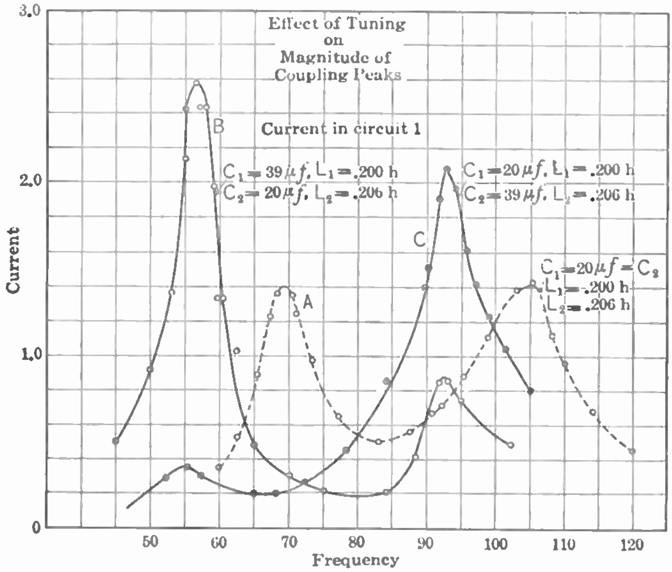


FIG. 54.

than R_2, but even with only 5 per cent coupling this will be true, for low-resistance coils.)

Increase the capacity of the primary by 50 per cent (leaving secondary at its previous value) and make another run. Then have the primary capacity at its original value and increase the secondary capacity by 50 per cent and make another run. Use tightest coupling for these two runs.

With frequency held at that value which gives resonance for a single circuit, and voltage held constant at about 150 per cent of that used in Exp. 6, make a run for various values of coupling, from the tightest available to the lowest value which can be used safely (voltage across primary condensers may get too high with very loose coupling).

EXPERIMENT 10

Object.—Use of alternating current bridge, at audio frequencies, to measure effective resistance, self and mutual induction. Measurement of resistance of a variable standard self-induction at various settings and frequencies, by resonance bridge.

Analysis.—The bridge is nothing but two paths connected in parallel to a common voltage source, as in Fig. 55. If then a connection is made to any point A in one path it must be possible to find a point in the other path such that there is no potential difference between these two points. If then they are connected together by some current-indicating device it will indicate zero, and the bridge is said to be balanced. Under such condition, by a simple analysis it is seen that $R_1/R_2 = R_3/R_4$. If say R_4 is being measured, R_3 will generally be a continuously variable resistance (or variable in very small steps), whereas R_1 and R_2 are known resistances adjustable in larger units. They are called the ratio arms of the bridge. In average laboratory procedure R_1, R_2, and R_3 will all be resistance boxes of the decade type. It is generally advisable to use for R_3 that box having the smallest units available. Thus if three boxes are available, two having 100's, 10's, and units and the third having 100's, 10's, 1's, and tenths, the latter should be used for R_3.

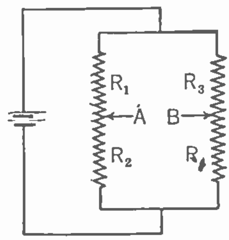


Fig. 55.

In the alternating current bridge, not only resistances (effective) are measured, but also inductances and capacities. The ratio arms are resistance boxes of the decade type, generally and the unknown inductance or capacity is balanced against a known variable inductance or capacity. Many types of alternating current bridges have been designed, but we shall discuss in this text only two or three of the more common ones.

The a.c. bridge is necessarily somewhat more difficult to use and analyze than the c.c. bridge, because as it is generally impossible to build devices having reactance which do not at the same time have resistance, It is thus necessary in order to get a balance to adjust both resistance and reactance.

This idea is illustrated in Fig. 56, in which the ratio arms are again shown as resistances R_1 and R_2, the unknown coil has reactance X_4 and effective resistance R_4 and the third arm, where the variable standard coil X_3 is placed, has also a variable resistance R_3. It is just possible that with this arrangement a balance cannot be obtained, no matter what ratio arms are used, nor what values of X_3 and R_3 are used. It is just possible that to get a balance there should be an additional resistance in series with the unknown coil, instead of in series with X_3, so that R_3 should be shifted to the arm containing X_4. To obviate the necessity of shifting this box it is advisable to use an additional box in the fourth arm. To indicate when a balance is obtained, it is generally convenient to use a head telephone as indicated in Fig. 57. This a.c.

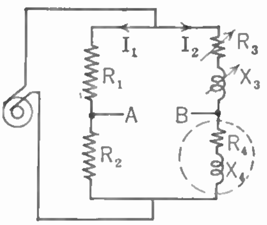


Fig. 56.

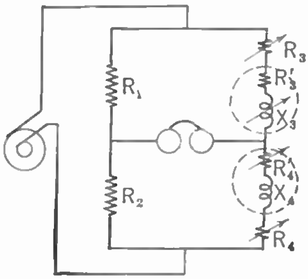


Fig. 57.

bridge is of great utility in the radio laboratory, and the student should gain as much ability in its use as possible.

The balancing of the bridge requires common sense as well as the ability to vary resistance boxes. When a balance is obtained we have

$$R_1/R_2 = (R_3 + R'_3)/(R_4 + R'_4) = X_3/X_4. \quad . \quad . \quad . \quad (34)$$

Thus the resistance must be balanced, as well as the reactance, if no sound is to be heard in the phones, and it is the necessity for the simultaneous balance of both quantities that bothers the beginner. However, by using a logical procedure it is possible to set up and balance the bridge in a few minutes.

In many college laboratories an entirely incorrect method is used in teaching the student to use this a.c. bridge. With a battery for power and c.c. galvanometer for detector, the bridge is balanced for resistance; the bridge is then transferred to an a.c. source of power and an a.c. indicator used in place of the galvanometer. Without disturbing the resistance elements of the bridge the reactance balance is obtained, as well as possible, by varying the standard reactance.

The reason why this procedure is not accurate is evident at once when we remember that the alternating current resistance of a coil is quite different from its value determined by continuous current measurement. The difference may be only a few per cent in some cases, but in others it may be several hundred per cent! And, of course, if the ratio arms and resistances R_3 and R_4 are left the same as they were adjusted for the continuous current balance, the alternating current test cannot possibly show a balance. It is true that as the variable standard inductance is varied, a minimum sound may be heard in the phone, but even with this minimum the bridge may be badly unbalanced.

There is no reason whatsoever for making the resistance measurement on continuous current test, because it is comparatively easy to balance the bridge for alternating current, when the right steps are followed. And the resistance value determined on c.c. test is not correct even if the bridge happens to show a complete balance when tested with alternating current, because *the resistance of the variable standard inductance may be quite different for the two cases.* In the case of the variable standards of many millihenrys, the thousand cycle resistance may be much more than it is for continuous current.

Before trying to balance the bridge on a.c. it is well to make some approximate estimates as to proper values of the ratio arms, as well as R_3 and R_4. With a little experience the inductance of the unknown inductance can be guessed, at least approximately. Thus if the coil is a single-layer solenoid, its inductance may be approximated if one or two somewhat similar solenoids of known values are compared with the unknown. For a given length of winding and number of turns, for example, this inductance varies about as the $(\text{diameter})^2$ of the solenoid. For a given size of wire and diameter of solenoid, the inductance increases nearly as the square of the length, for short coils, and nearly as the length, to the first power, for long coils. For a given length and diameter of coil the inductance varies as the square of the number of turns per inch, etc. From ideas of this kind the inductance may be guessed at, and the ratio arms chosen so that the mid-value of the variable standard inductance will balance the bridge.

The resistance of a solenoidal coil, at 1000 cycles, may be taken as about one ohm per millihenry, for those sizes of wire generally used in radio coils. So if the resistance of the variable standard is known even the values of R_3 and R_4 may be approximately set, after the required ratio (to balance the inductance) has been set. Then with about one volt impressed on the bridge, the variable standard inductance is varied throughout its range. A minimum of sound in the telephone receiver should be heard somewhere on the scale. With the standard left at this setting, one of the resistances R_3 or R_4 (say R_3) is increased; if the phone signal decreases, R_3 is further increased until a further increase increases the sound. If upon increasing R_3 this signal increases, it must be decreased, or, if it is already set at zero, R_4 must be increased to reduce the signal to a minimum. When this minimum has been reached, the variable standard inductance is reset for a new minimum, the resistance R_3 or R_4 is further adjusted to reduce the signal, and these two approximations at a balance will generally give a balance so close as to make the signal inaudible. Then eq. (34) may be used to calculate the unknown inductance. To calculate the resistance of the inductance being measured, it is necessary to know the resistance of the standard inductance for the setting and frequency being employed, if this is not known it must be measured as explained later in this test.

If when first trying for a balance, by varying the standard inductance, a minimum of phone signal is heard at one end or the other of its scale setting, the ratio of the bridge must be changed. Thus referring to Fig. 57, if minimum signal is heard at the minimum value of X_3 (the standard) R_1 must be increased or R_2 decreased. This change must be continued until an approximate balance is obtained (as X_3 is varied) somewhere on the scale of X_3 not too close to either limit of its adjustment. Theoretically the impedance of all arms of the bridge should be the same, to bring about the most sensitive condition for balancing. But actually it is possible to balance a bridge of this kind, as accurately as the scale of the variable standard can be read, with the R_3, R_4 arms ten or even a hundred times as great (or small) as the R_1, R_2 arms. It is also possible to balance the bridge with the ratio widely different from unity, but certain errors are likely to occur which make it inadvisable to use a ratio greater than perhaps five or ten.

It will be realized that it is not advisable to try and measure the inductance of a coil of a few microhenrys by this method; the reactance of such a coil is so low as to make a balance very uncertain. At frequencies in the audible range, coils of a millihenry may reasonably well be determined, but the bridge, as ordinarily arranged, should have inductances of several millihenrys up to a henry, for easy and accurate determination.

If great accuracy in balancing is desired it is advisable to use one or two stages of audio frequency amplification between the bridge and detector; this is especially advisable if the laboratory is noisy.

It will generally be found that, with the bridge as described so far, it is impossible to adjust conditions to give zero noise in the phones; a minimum can be reached but not zero. This is frequently due to the fact that even when the bridge is perfectly balanced, there is still current running from the bridge into the phones, thus giving a signal. The observer is generally near to ground potential, and the parts of the bridge where the phones are attached may not be at ground potential (they generally will not be so). Thus charging current flows through the phones into the observer (by capacity action) and so to ground. This current is what gives the residual noise, when the bridge is perfectly balanced.

By using what is called a Wagner earth connection, this difficulty can be overcome. As shown in Fig. 58, another auxiliary bridge is added to the one we already have, consisting of the two resistances R_5 and R_6 (of total resistance, say, 1000 ohms), the junction of which is earthed. An extra pair of phones T_2 is used (the pair T_1 can be temporarily transferred) to balance the bridge made up of the original bridge, in combination with R_5 and R_6.

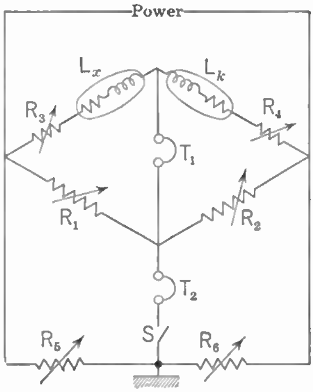


FIG. 58.

With the switch S open the original bridge is balanced as closely as possible. Then switch S is closed, T_2 used for a detector, and the extra bridge is balanced by varying R_5 and R_6, leaving the original bridge as balanced before. When this "earth balance" has been obtained, S is opened, and the final adjustment of the original bridge is carried out.

For the measurement made at a fixed frequency a hummer generator may be used for power; for the resistance measurements calling for various frequencies, either an audio frequency oscillator or a beat frequency oscillator may be used. Of course, frequency calibration curves for these are also necessary.

For inductances measured in microhenrys a higher frequency power supply should preferably be used; this is reasonable when it is remembered that such low inductance coils are never used in audio frequency circuits. Their reactance, at audio frequencies, is too low to make them of service.

For inductances measured in many henrys the bridge is satisfactory when a sufficiently large variable inductance is available. There are, however, other methods (see Exp. 36) which serve without the use of such a large standard inductance, which is generally not available.

To measure the mutual induction of two coils they are connected in series, after being placed in the position for which the mutual induction is desired. The total self-induction of the pair is measured. Then the connection of one of the coils is reversed and another value of total self-induction is obtained. (The resistance balance will probably be different when the connection has been reversed, so that R_3 or R_4 must be charged. Why?) One-quarter of the difference of the two self-inductions so measured is the value of the mutual induction desired.

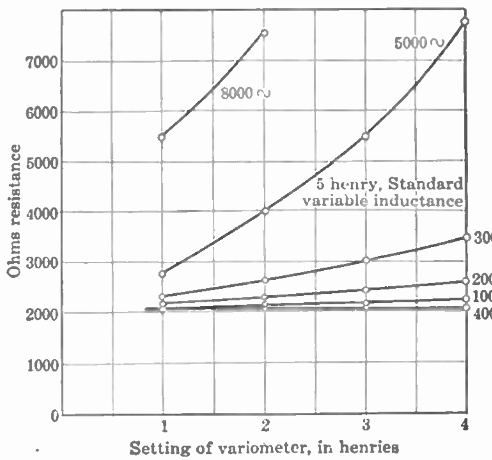


FIG. 59.

As mentioned before, to obtain the resistance of the coil measured in the test it is necessary to know the resistance of the standard. As this varies with frequency, *and with setting*, it is necessary to have a suitable set of curves to give the resistance for any setting and frequency. Typical values for two variable standard inductances built by a well-known manufacturer are shown in Figs. 59 and 60; evidently standards of the higher values of self-induction increase their resistance at the higher frequencies faster than those of lower self-induction.

To measure the resistance of the standard for various positions and frequencies, an adjustable condenser of good grade is necessary; this is connected in series with the standard inductance in one arm of the bridge, the other three arms being resistances only. By selecting the right value of capacity in the adjustable condenser the inductive reactance of the standard is neutralized, thus giving in this arm also only resistance.

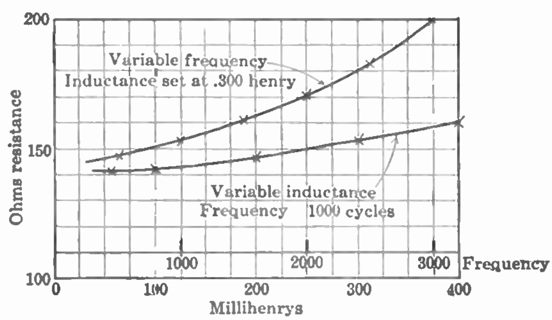


FIG. 60.

When a balance is obtained the resistance of the standard is at once obtained (approximately) because the effective series resistance of a good condenser is negligible compared to that of the inductance, so that all the resistance measured may be considered as coil resistance. If, however, it is desired to have the resistance of the standard inductance as accurate as possible, it may be assumed that the power factor of a good paraffin paper condenser is ½ per cent. (This is about as good as they can be built.) It may be assumed that this power factor is independent of frequency, throughout the audio frequency range. In case a mica condenser is used, its power factor may be assumed as 0.2 per cent, and in case a good air condenser (in which no solid dielectric is used) is employed, the power factor may be assumed, without very much error, as 0.1 per cent.

Suppose the 50 millihenry value of a variable standard is being measured, at 1000 cycles. Its reactance is 314 ohms, so the condenser used to give resonance will also have 314 ohms. This requires a capacity of about 0.5 microfarad. Probably such a condenser would be of the paraffin-paper type, and its resistance would be ½ per cent of 314 or 1.5 ohms. The power factor of the ordinary variable standard inductance, at 1000 cycles, is about 0.2, so its resistance, for the point being measured, would be found by the bridge balance to be, let us say, 60 ohms. From this 60 ohms (which represents the resistance of the inductance and condenser in series with each other), we subtract 1.5 ohms as the probable series resistance of the condenser, thus obtaining a value of 58.5 ohms for the inductance.

To measure the resistance of the variable inductance throughout its range, and throughout the audio frequency range where it is likely to be used, requires quite a range in capacity, as we will now show. Ordinarily a range of 10 to 1 is obtained in a variable inductance, thus the one we have in mind might have a range of 15 to 150 millihenrys. The lowest audio frequency at which we wish to use it, let us say, is 200 cycles, and the highest 5000 cycles. We will then calculate how much capacity it requires to balance 15 millihenrys at 200 cycles and how much for 150 millihenrys at 6000 cycles, and these will be the extreme values. For the first condition we require 43 microfarads, and for the last condition about 0.007 microfarad. To obtain this range of values we require some trays of telephone condensers, a decade box of capacity units, having ten steps each of 0.1, 0.01, and 0.001 microfarad, and a variable air condenser, having a maximum capacity of about 0.0015 microfarad. These are readily obtainable on the market.

For each frequency, say 200, 500, 1000, 2000 and 5000, about 5 points are to be obtained throughout the inductance range. For the low frequencies the variable air condenser will not be necessary, the telephone condensers in parallel with the decade condenser box being sufficient for a good balance. For the higher frequencies the decade box and the variable air condenser, in parallel, will be required.

In performing this part of the test the student will be much bothered, at first, by the harmonic voltage of the power supply and will probably balance the bridge very poorly. All the readily available sources of power supply for this test (a tuning-fork oscillator, or a vacuum tube oscillator) generate quite appreciable harmonics; in many cases the third harmonic is as great as 10 per cent of the fundamental.

Now when the right capacity has been found to neutralize the reactance of the inductance, and the resistance balance has been obtained, both for the fundamental frequency, neither condition is satisfied for the harmonics, so they will not be balanced and so will produce a loud signal in the telephone receivers. Moreover, if the bridge is being balanced for 200 cycles, let us say, the third harmonic will be 600 cycles and the fifth harmonic will be 1000 cycles, and both the ear and phones are much more sensitive for these frequencies than they are for the fundamental, 200 cycles. Hence even when the balance is actually obtained (for the 200 cycles) the noise in the telephone receiver may be almost as great as before the bridge was balanced.

The experimenter must learn to listen only to that frequency with which he is working, and this is one of the valuable lessons to be learned from this test. To gain experience in this it is well to listen to a bridge, balanced for the fundamental, as the variable inductance is varied a little either side of the point for which the balance has been obtained. The loud harmonic tones do not change in intensity as this is done, but the fundamental tone can be heard underneath the overtones, waxing and waning as the balance is destroyed or obtained.

Check the laboratory variable standard inductance against one or more fixed standard inductances. Try this for several different bridge ratios, to see if the value determined is independent of bridge ratio.

Measure several single-layer solenoids, of one millihenry or more. Get dimensions of the coils for calculating the inductance by Nagaoka's formula.* Get the inductance of a two-layer solenoid, one layer being wound on top of the other. Preferably this should be the duplicate in dimensions and size of wire, of one of the single-layer solenoid measured. Its inductance should be found to be about four times that of the single-layer solenoid. Measure the inductance of a head telephone, both by comparison with a variable standard and by finding the capacity required for resonance, the frequency, of course, being known. Measure the

* Principles of Radio Communication, 2d ed., p. 194.

mutual inductance of a pair of coils throughout the range of their possible motion. This may well be a pair of coils which are to be used later in the course, for coupling the grid and plate circuit of a triode for producing oscillations; the calibration will then be needed. Measure the resistance of the variable standard inductance throughout the ranges specified by the instructor.

EXPERIMENT 11

Object.—Use of special bridge for measuring capacity at audio frequencies. To measure phase angle of a condenser. Calibration of a variable condenser of the SLC, SLF, and the logarithmic type. Measurement of specific inductive capacity and " phase difference " of various dielectrics.

Analysis.—In the previous experiment the impedances used in the arms of the bridge are comparatively low, very seldom over 1000 ohms, and generally much less than this. Thus a one millihenry coil, for use in a radio circuit, tested at 1000 cycles, has only 6 ohms reactance and a few ohms resistance. When such low impedances are used in the bridge arms, it will be found that inappreciable errors are incurred, owing to capacities to ground, extraneous induced voltages by either electric or magnetic fields, etc. Thus the bridge may be made up of several decade resistance boxes suitably connected together, and it will be found that if reasonable neatness is used in the arrangement and wiring, this bridge can be torn down and reconnected time after time, and the measured values of inductance and resistance will be the same time after time.

Now when we endeavor to measure, at audio frequencies, the condensers used in radio circuits, we find that great difficulty is encountered in getting results at all consistent, and a little reflection at once brings out the difference between this case and that of the previous test.

The ordinary tuning condenser used in a radio set has a maximum capacity of about 300 micro-microfarads. At a frequency of 1000 cycles this condenser has a reactance of 500,000 ohms! The slightest unbalance of a bridge, made up of such elements, with respect to grounding, stray capacities, stray induced voltages, etc., will give large errors in the measured values of the capacities and resistances. In making measurements of this kind only bridges of very special construction can be used, if the results are to be at all accurate.

Errors due to stray capacities to ground are least when the ratio arms of the bridge have the same resistance, so we notice first that bridges intended for this kind of work have unity ratio; the ratio arms are set at some reasonable value and are not adjustable. Although theoretically the ratio arms should have the same impedance as the arm being measured, and this is generally hundreds of thousands of ohms in work of this kind, it is found advisable to make the ratio arms of about 5000 ohms resistance only. Mechanical difficulties, lack of permanence in resistance, etc., are encountered if the wire used in making these resistance arms is too fine. And if a very large resistance is attempted using reasonably sized wire, the size of the resistance unit is so large that difficulties from its distributed capacity are encountered. So the compromise value of about 5000 ohms is found best.

To eliminate the possibility of voltages being induced in one part of the bridge by other parts, it is necessary to shield the various parts carefully. This is done by putting the various parts in separate copper-lined partitions and grounding these partitions. Such a copper lining is a perfect shield for electrical fields, and a reasonably good shield for magnetic fields, of audio frequencies.

The variable standard condenser which necessarily is used with such a bridge, we will analyze later. It is generally necessary to have a variable resistance in series with one condenser or the other, if a perfect balance is to be obtained, so there is a decade resistance box built into the bridge, as well as the ratio arms.

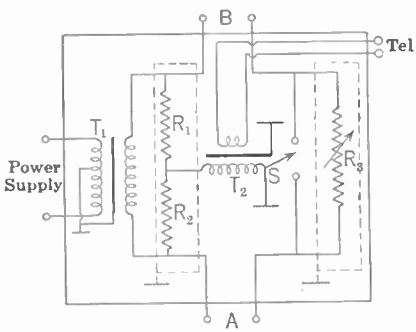


FIG. 61.

The power supply for the bridge may greatly unbalance the bridge, with respect to ground, so that care must be exercised here. The bridge we are describing is illustrated in Fig. 61, and it will be noticed that the power supply is connected to the bridge through a transformer and that there is a grounded metallic shield between its two windings. This isolates the bridge from the power supply circuit, in so far as electric fields are concerned.

The observer, on account of his proximity with the bridge circuit through the head phones, may produce leakage capacity effects which seriously affect the accuracy of such a bridge; the phones are therefore connected to the bridge through another transformer, and this also has a grounding shield between the two windings. This transformer is generally not unity ratio but "steps down" from bridge to phones in the ratio of about three to one. This makes the phones more nearly match the impedance of the bridge arms.

It will be noticed that the two resistance arms of the bridge are in the same copper-lined partition; it has been found unnecessary to shield them from each other, when they are mechanically and electrically similar.

It will be noticed from Fig. 61 that the variable resistance R_3, in a separate copper-lined section of the bridge, may be thrown from one side of the bridge to the other by the switch S. The two condensers to be compared, the standard and the unknown, are connected to the terminals A and B, and it will be seen that by throwing S down the variable resistance is put in series with the condenser connected to B, and vice versa.

The bridge network itself is grounded at the switch S. It looks as though it would be just as well to ground the bridge either at this point or at the junction points of the two ratio arms; owing to the action of stray capacities, however, it is found that large errors may occur in the resistance determination of the unknown condenser if the bridge is grounded at the resistance arms. This may amount to 50 per cent or more if the condenser being measured has a low capacity.

The standard condenser used with such a bridge must be of excellent mechanical construction, as well as electrical, or the bridge is useless. It must be possible to adjust the condenser, easily, by amounts measured in hundred-thousandths of its full scale capacity if resistance determinations are to be made with any degree of precision, as we will now show.

A good variable condenser has an equivalent series resistance not more than .01 per cent of its reactance. Thus if the condenser has a reactance of 500,000 ohms, its series resistance may be 50 ohms, and we may want to know this resistance, to within 5 ohms, let us say. Now the a.c. bridge is an *impedance* bridge; when it is balanced the *impedance* in the arms are balanced, and if we wish to obtain the resistance within 5 ohms we must balance the impedance within 5 ohms. (This is an approximate statement; for the skilled observer it is not quite true.) But the impedance is practically all reactance, so we may say that the reactance must be balanced to within 5 ohms. But this is only 0.001 per cent of the reactance, so it will be easily appreciated that the standard condenser must be adjustable by the most minute steps.

Now a change in capacity of much more than 0.001 per cent will occur if the hand is brought in the vicinity of the condenser, and the hand must necessarily be brought somewhere near the condenser to adjust it. It is thus evident that the standard condenser must be electrically shielded, so that the proximity of conducting, or semiconducting, bodies does not affect the capacity.

A good standard condenser therefore is mounted in a box which is copper-lined, and this copper lining is well connected to the *rotating set* of plates. It is easier to isolate thoroughly the stationary plates than the rotating ones, so the stationary plates form the ungrounded side of the condenser (sometimes referred to as the high potential side or merely the "high side" of the condenser). In connecting such a condenser to the bridge of Fig. 61 it is imperative that the grounded plates connect to the posts going to switch S; if it is connected in the reverse manner very large errors, especially in resistance determinations, will be incurred.

In measuring the capacity of an unknown condenser, therefore, it is connected to either A or B terminal of the bridge, and the standard is properly connected to the other. As the bridge has unity ratio it follows that a balance can be obtained only if the unknown has a capacity within the range of the standard.

With the resistance R_3 set at zero (switch S thrown to either side) the standard condenser is varied until a reasonable balance is indicated. Some resistance is then cut in on R_3 and its effect on the balance is noted. (Unless the capacity balance has been quite carefully obtained probably no effect will be noticed from quite large changes in R_3.) If the balance is improved R_3 is changed until a new minimum is given by the phones; if the balance is made poorer by increasing R_3 from its original zero value, switch S should be thrown to the other side and the above adjustment carried out. After R_3 has been set at its best value the standard condenser is adjusted for a new minimum and then R_3 again adjusted. When the best balance obtainable is reached, the reading of the standard and of R_3 is noted. The connections of the standard and unknown to the bridge are then interchanged, and it will generally be found that the other values of capacity and resistance are required for balance. The average of the two sets of values should be used as the correct one.

The reading of the standard condenser gives at once the capacity of the unknown, and the value of R_3 enables us to get the resistance of the unknown in terms of the resistance of the standard condenser. Generally it will be found that R_3 must be connected in series with the standard to get a balance, because it is very seldom that a commercial condenser is built with losses smaller than those of the well-built variable standards.

Suppose the average of the two values of R_3 required for a balance is 85 ohms in series with the unknown. This means that the unknown has 85 ohms more resistance than has the standard, so we must know the resistance of the standard. This generally requires the use of a formula the basis for which the student must appreciate.

There is no loss at all in the air between the stationary and rotating plates, but there is a dielectric loss in the blocks of insulation which serve

Dielectric Constant, Phase Difference and their Product for Several Commercial Insulating Materials*

Material	Frequency kilocycles	Dielectric constant	Phase Difference, degrees	Product
Phenol Fiber A	295	5.9	2.9	17.1
	500	5.8	2.9	16.8
	670	5.7	2.9	16.5
	1040	5.6	3.3	18.5
Phenol Fiber B	190	5.8	2.2	12.7
	500	5.6	2.5	14.0
	675	5.6	2.6	14.6
	975	5.6	2.8	15.7
Phenol Fiber C	200	5.4	2.1	11.3
	395	5.4	2.2	11.8
	685	5.3	2.3	12.2
	975	5.2	2.4	12.5
Phenol Fiber D	194	5.4	4.2	22.7
	500	5.2	3.9	20.3
	695	5.2	3.9	20.3
	1000	5.1	3.8	19.4
Wood (Oak)	300	3.2	2.1	6.7
	425	3.3	2.0	6.6
	635	3.3	2.2	7.3
	1060	3.3	2.4	7.9
Wood (Maple)	500	4.4	1.9	8.4
Wood (Birch)	500	5.2	3.7	19.2
Hard Rubber	210	3.0	0.5	1.5
	440	3.0	0.5	1.5
	710	3.0	0.5	1.5
	1126	3.0	0.6	1.8
Flint Glass	500	7.0	0.24	1.68
	720	7.0	0.24	1.68
	890	7.0	0.23	1.61
Plate Glass	500	6.8	0.4	2.7
Cobalt Glass	500	7.3	0.4	2.9
Pyrex Glass	500	4.9	0.24	1.18

* All of the samples had been in the laboratory for some time during summer weather without artificial drying or other special preparation.

as mechanical supports to keep the two sets of plates in their proper relative positions. Hard rubber, glass, isolantite, or quartz are used for these blocks; all of these have small losses and are mechanically suitable materials to use, especially the two last mentioned. The characteristics of some commercial dielectrics and their variation with temperature are shown in the accompanying tables taken from the Bell System Technical Journal for Nov. 1922.

VARIATION WITH TEMPERATURE OF DIELECTRIC CONSTANT, PHASE DIFFERENCE AND THEIR PRODUCT FOR SOME COMMERCIAL INSULATING MATERIALS (FREQUENCY 500 KILOCYCLES)*

Material	Temperature, degrees C.	Dielectric constant	Phase difference, degrees	Product
Molded Phenol Product A	21	5.6	3.1	17.4
	71	6.9	6.5	45.0
	120	10.4	22.0	230.0
Molded Phenol Product B	21	5.2	2.3	12.0
	71	6.1	3.7	22.5
	120	7.6	8.9	68.0
Molded Phenol Product C	21	5.3	2.8	14.8
	71	6.1	3.6	22.0
	120	6.7	9.6	64.0
Phenol Fiber B	21	5.6	2.5	14.0
	71	6.6	3.1	20.5
	120	6.5	4.6	30.0
Phenol Fiber C	21	5.4	2.3	12.4
	71	6.0	3.9	23.5
	120	5.3	4.9	26.5
Phenol Fiber D	21	5.2	3.9	20.3
	71	6.6	6.9	46.0
	120	6.3	12.5	85.5
Hard Rubber	21	3.0	0.5	1.5
	71	3.1	1.2	3.7
	120	3.2	3.7	11.8
Pyrex Glass	20	4.9	0.24	1.18
	74	5.0	0.4	2.0
	125	5.0	0.7	3.5

* The measurements on each sample were made in the order in which they are given in the table.

The actual condenser may then be represented as shown in Fig. 62. C represents the capacity between the rotating and stationary plates through the air dielectric, and C_0 represents the capacity between the two sets of plates where they are clamped together by means of the insulating blocks. The dielectric, having losses, is shown by the sectioned part of the sketch. The value of C_0 is very small compared to the maximum value of C, being perhaps 1 per cent the maximum capacity of the condenser. The arrangement of Fig. 62*a* can be changed to its equivalent Fig. 62*b*, in which the losses of the faulty dielectric are now replaced by the losses in the resistance R_0. The power factor of such material as is used in a good condenser may be 0.005, so that if C_0 is, say, 15 $\mu\mu f$, then at 1000 cycles R_0 must be $0.005 \times \frac{10^{12}}{2\pi \times 10^3 \times 15}$, or about 5000 ohms.

Thus the standard condenser we have in mind is equivalent to the

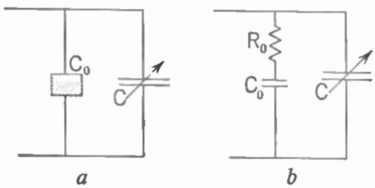


FIG. 62.

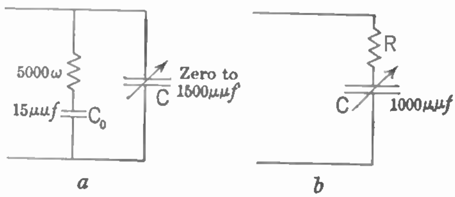


FIG. 63.

arrangement of Fig. 63*a*. If now we wish to replace the arrangement of Fig. 63*a* by that of Fig. 63*b*, how large must R be? It must have the same losses as those occurring in R_0, and as these losses are given by I^2R, the I^2R loss in R (Fig. 63*b*) must equal that in R_0 (Fig. 63*a*).

But for a given frequency and voltage the current into a condenser varies as the capacity, so the current flowing in R is equal to C/C_0 times that flowing in R_0. So that R must be given by the value of $R_0(C_0/C)^2$. If we assume that C is $1000\mu\mu f$ then we find R to be 10 ohms, so that our standard condenser is representable at 1000 cycles and a setting of $1000\mu\mu f$ by this value of capacity in series with 10 ohms resistance.

For other frequencies and condenser settings the equivalent series resistance must be suitably changed. *The equivalent series resistance varies inversely with the frequency and inversely with the square of the capacity setting.* Thus if the standard has its capacity increased to $1500\mu\mu f$ the current through R (Fig. 63*a*) is 1.5 times as large as it was for a setting of $1000\mu\mu f$. But the losses in the insulating blocks is just the same as it was before, so the losses in the new value of R must be the same as they were before or $R_{1500} = R_{1000} \times \left(\frac{1}{1.5}\right)^2$, or about 4 ohms. On the other hand, if C is decreased to a value of $100\mu\mu f$, the value of R will be 100 times the value for $1000\mu\mu f$ or 1000 ohms.

For a frequency of 500 cycles, and a capacity setting of $1000\mu\mu f$, the equivalent series resistance is 20 ohms.

If then the manufacturer of the standard condenser gives the value of the equivalent series resistance for one frequency and setting, its value for other frequencies and settings may be calculated.

All of the foregoing argument is based on the assumption that the insulation resistance between the plates of the condenser is so high that the loss due to leakage current is negligible compared to that due to dielectric hysteresis in the material of the separating blocks of insulating material. It may well be that such is not the fact.

Let us suppose that the insulation resistance is equal to 1000 megohms, a fairly high resistance. This shunt resistance may be changed to its equivalent series resistance by eq. (23), page 55.

$$R_s = \frac{X^2}{R_{sh}}.$$

For $1000\mu\mu f$ at 1000 cycles the value of X is 160,000 ohms, so we have $R_s = \frac{2.5 \times 10^{10}}{10^9} = 25$ ohms! If then sufficient moisture condenses on the separating blocks to reduce its insulation resistance from let us say 100,000 megohms to 1000 megohms, the equivalent series resistance will be increased from 10 ohms to 35 ohms! This is not at all an impossible condition in a laboratory during the humid summer days.

The losses due to leakage over the surface of the dielectric do not follow the same laws as do the dielectric losses, so that this variation in equivalent series resistance deduced for the latter case will not suffice if the leakage loss is appreciable. The equivalent series resistance, in so far as it is due to leakage current through, or over the surface of, the insulating blocks, varies inversely with the square of the condenser setting and inversely with the square of the frequency.

It is to be noticed also that the block of insulation, on the bridge itself, to the terminals on which the standard condenser is connected, is just as important in this discussion as the insulating block of the standard condenser itself; this block is connected directly in parallel with the standard condenser, so that its characteristics function as though it were part of the standard condenser itself.

From the foregoing discussion, it is evident that even when this special capacity bridge has been balanced, the resistance of the condenser under test, deduced from bridge readings and manufacturers' data, may be very far from the correct value.

There is another method (the differential method) of using this capacity bridge, by which practically all of the previously discussed possible errors may be eliminated, and this method, now to be described, is the normal method of using such a bridge. It is suitable for measuring condensers of capacity somewhat less than the capacity of the standard; say with a standard of $1500\mu\mu f$ (an ordinary value), a condenser of $1400\mu\mu f$ can readily be measured.

The standard is properly connected to the bridge (proper polarity), and a variable condenser of equal range (or a fixed condenser of value practically the full range of the standard) is connected to the other side of the bridge. The condenser to be tested is connected in parallel with the standard with short, rigid wires. In case the condenser is of the shielded type, its shielded side must be connected to the corresponding side of the standard.

The wire connecting the "high side" of the standard to that of the unknown is now disconnected, *at the unknown condenser*, bending the wire enough to put it about $\frac{1}{4}$ inch away from the terminal on the unknown condenser.

With the standard condenser set at practically full scale the variable condenser in the other arm is varied until a balance is obtained. Generally it will be necessary to have R_3 in series with the standard in getting this balance. Let the reading of the standard condenser and resistance R_3 be called C_1 and R_1, respectively.

Now the unknown condenser is connected in circuit and the bridge again balanced by varying the standard condenser and R_3. Call the new reading C_2 and R_2.

Then the capacity of the unknown condenser is at once given by $C_1 - C_2$ and its resistance is given by the equation

$$R = (R_1 - R_2)\left(\frac{C_1}{C_1 - C_2}\right)^2$$

as proved below.

In case it is necessary to shift R_3 from one arm to the other to get this balance, the sum of its two readings must be taken rather than the difference.

In Fig. 64*a* is shown the standard condenser C, its equivalent series resistance R_a for the setting, and R_1, the required value of the R_3 resistance of the bridge to get a balance.

When the unknown condenser has been connected in parallel with the standard and the balance restored by capacity and resistance adjustment, the circuit is as shown in Fig. 64*b*. The reactance of these condenser circuits is so much greater than the resistance that we may consider reactance and impedance equal to each other. Then

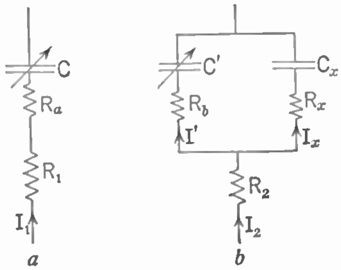


FIG. 64.

$$C = C' + C_x \text{ or } C_x = C - C'.$$

If the two circuits above are equal in resistance they will have the same loss when a given voltage E is impressed across them, so

$$I_1^2(R_1 + R_a) = I_2^2R_2 + (I')^2R_b + I_x^2R_x$$

and

$$I_1 = \omega CE,\ I_2 = \omega CE,\ I' = \omega C'E,\ I_x = \omega C_xE$$

so

$$\overline{\omega C}^2(R_1 + R_a) = \overline{\omega C}^2R_2 + \overline{\omega C'}^2R_b + \overline{\omega C_x}^2R_x$$

Now divide through by ω^2 and note that, as the resistance of a condenser varies inversely with the square of its capacity,

$$C^2R_a = C'^2R_b,$$

and we get

$$C^2R_1 = C^2R_2 + C_x^2R_x$$

or

$$R_x = (R_1 - R_2)\left(\frac{C}{C_x}\right)^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (35)$$

In this method of measuring the capacity and resistance of a condenser, stray capacities, losses in terminal blocks of the bridge, etc., do not enter into the measurement; the only assumption made is that the equivalent series resistance of the standard varies inversely as the square of its capacity.

To balance the resistance of the bridge (R_3) to one ohm, when the capacity (standard plus unknown) is $1000\mu\mu f$, requires the capacity balance to be obtained to $.01\mu\mu f$ and necessitates generally a two-stage amplifier between the detector and bridge and about 50 volts impressed on the bridge. It is thus evident that very accurate determination of the equivalent series resistance of a condenser, by bridge method, is not easy.

In measuring the specific inductive capacity and phase angle of a dielectric, it is advisable to have the material to be tested in the form of thin slabs, about 10 inches square. Metal plates are clamped on opposite sides of the slab, and just opposite to one another. If the slab is not reasonably smooth, somewhat more accurate results will be obtained if vaseline is smeared over the slab, before the metal plates are clamped on it. To make the " edge effect " negligible the plates should be at least 4 inches in diameter. It is well to " back up " the metal plates with thick wooden blocks to make them lie close to the surface of the dielectric. A small air gap between the dielectric sample and metal plate will appreciably reduce the capacity of the combination.

By measuring the capacity of the condenser so formed, and obtaining its dimensions by caliper and rule, the specific inductive capacity can be calculated, using the relation $C = \frac{AK}{4\pi d}$. The capacity is given in centimeters if the plate area A, and dielectric thickness d, are given in centimeters. The capacity in micro-microfarads is obtained by dividing the capacity in centimeters by 0.9.

In a perfect condenser the current leads the voltage by 90°; in actual condensers the lead is always something less than this. The difference between the angle of lead of the perfect condenser (90°) and the actual lead is frequently called the " phase difference " of the condenser.

As the series resistance of a condenser is always small compared to its reactance, the reactance and impedance may be considered equal and the phase difference is then given by the expression

$$\psi = R\omega C \text{ radians} = 57.3\, R\omega C \text{ degrees} \quad . \quad . \quad . \quad . \quad (36)$$

By the first method described (standard on one side of the bridge, unknown condenser on the other) measure the capacity and resistance of two or three laboratory condensers, paraffin paper and mica. After each determination reverse the position of standard and unknown and make a check determination.

Using the differential method calibrate and measure the resistance of three laboratory condensers of different types, say a straight line capacity, a logarithmic capacity, and a straight line frequency condenser.

Using the differential method measure the specific inductive capacity and phase difference of several samples of dielectrics.

EXPERIMENT 12

Object.—Measurement of the internal capacity of various types of vacuum tubes.

Analysis.—The vacuum tube used so much in radio has at least three electrodes, and sometimes four or five, small metallic surfaces which necessarily act like small condensers. In the three-electrode tube, having cathode, grid, and anode, the capacity between the grid and anode plays a very important role in its operation; the other two capacities are of somewhat less importance.

These small electrodes, separated by vacuum for dieletric, must obviously form condensers of extremely small capacities; in the average detector and amplifying tubes they have capacities ranging from 1 to 10$\mu\mu f$, seldom more. Of course in the larger tubes used to obtain many watts of power output, the capacities are somewhat larger.

Not only do the electrodes themselves contribute to this capacity, but the connecting wires (especially where they go through the " press ") and the base with its prongs also add an appreciable amount.

In the building of radio frequency amplifiers these small condensers play an extremely important role, and it is desirable to know them to one-tenth of one micro-microfarad. Evidently such a measurement is not readily obtainable with the bridges so far discussed; because of this a special bridge has been designed and is used for this specific use.

At 1000 cycles the reactance of a 5$\mu\mu f$ condenser is 30,000,000 ohms; with such impedance in the arms it is evident that stray capacities to ground, etc., would entirely nullify the value of any measured values, unless the scheme of measurement was free from such effects. As pointed out in the last experiment, the differential method of capacity measurement, properly used, is independent of ground capacities, stray induced voltages, etc., and this is therefore the scheme generally used to measure the inter-electrode capacities.

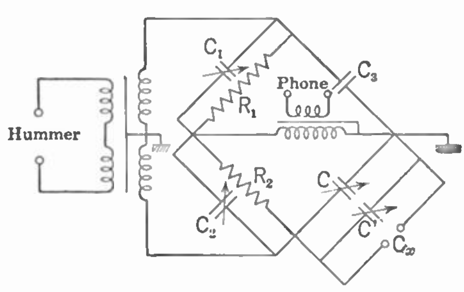


Fig. 65.

The arrangement of apparatus is as shown in Fig. 65. A hummer (generally self-contained) supplies power to the bridge through a transformer, which is built in two equal sections, grounded at the junction of the secondaries, and electrostatically shielded between each primary and secondary winding. The two resistance arms R_1 and R_2 are from 100,000 to 500,000 ohms each; they are not adjustable, but must be reasonably equal in resistance. Thus the bridge will operate satisfactorily if one is 450,000 and the other 510,000 ohms, but, as put on the market by the manufacturer, it would not operate if one resistance is 25 per cent greater than the other. The reason for this will appear later.

A fixed condenser C_3 has about $30\mu\mu f$ of capacity and forms the third arm of the bridge. The fourth arm is made up of a standard variable condenser of about $30\mu\mu f$ capacity, C, a small uncalibrated variable vernier condenser C' in parallel with C, and the pair of terminals for connecting in the unknown condenser C_x. The telephone is connected to the bridge through a step-down transformer, to make its impedance match the bridge arm impedances. Across the resistance arms R_1 and R_2 is connected a special, small, variable condenser consisting of two sets of stator plates, and one set of semicircular rotor plates. These rotor plates mesh with either set of stator plates, or partly with one and partly with the other. This type of condenser acts just like two condensers, so coupled together mechanically that as the capacity in one increases the other correspondingly decreases. This compound condenser is called the "loss adjuster"; its function and operation will be described later.

The standard condenser C is a straight line capacity condenser, having a constant increment in capacity per degree of rotation. This calibration factor, i.e., capacity per scale division, is furnished by the manufacturer, and is generally about $0.3\mu\mu f$. The scale of C is mounted on the shaft in such a way that when it reads zero the capacity is at its maximum, instead of minimum, value.

In using the bridge C is set at zero and the vernier C' is adjusted to make the bridge balance. It generally happens that the losses in the C_3 arm and the C arm are not the same, and so the bridge can be only approximately balanced. But if the power factors of the C_3 and C arms are not the same, the bridge still can be balanced if the power factors of the R_1 and R_2 arms are suitably altered. This is the function of the C_1–C_2 condenser. When the bridge has been balanced as closely as possible by C', the C_1–C_2 condenser is adjusted to give a better balance; C' can then be further adjusted for a new minimum.

Now the unknown condenser is connected at C_x and of course the balance is destroyed. C is now reduced (its scale reading increasing from zero) so as to restore the balance as well as possible. The loss-adjusting condenser C_1–C_2 will probably require a slight change, then C is again set to get as good a balance as possible. It is possible that C_1–C_2 must again be changed to get a balance as good as desired, but not likely. At any rate C and C_1–C_2 are alternately adjusted to get the best balance possible.

The reading of the scale of C, multiplied by the calibration factor, then gives the value of C_x, the unknown condenser. This can generally be found to about $0.01 \mu\mu f$. There are many other schemes available in the well-equipped radio laboratory for measuring the inter-electrode capacity of tubes, but none as convenient or simple as this one.

Now it so happens, when it is attempted to measure the internal capacities of the triode by this method, that although very accurate measurements can be made, the value of capacity obtained from the bridge balance is not really the capacity we have been attempting to measure, but a more complex quantity.

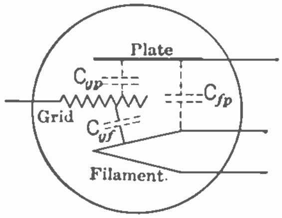


Fig. 66.

In Fig. 66 is a conventional sketch of a triode; indicated in the sketch are the three capacities mentioned, that between plate and filament, C_{pf}; that between plate and grid, C_{pg}; and that between grid and filament, C_{gf}.

Now when we measure the capacity between the plate and filament terminals, it is not merely C_{pf} that we measure, but this condenser plus a parallel capacity made up of C_{pg} and C_{gf} in series.

Now from the law of series condensers we know that the capacity of condensers C_{pg} and C_{gf} in series is equal to $\frac{C_{pg}C_{gf}}{C_{pg} + C_{gf}}$, so that the capacity actually measured between the plate and filament terminals is equal to

$$C'_{pf} = C_{pf} + \frac{C_{pg}C_{gf}}{C_{pg} + C_{gf}}. \quad \quad (37)$$

A similar relation holds good when the capacity is measured between any other pair of terminals, so that all of the capacities are measured too large.

Of course, having measured the three apparent capacities, and knowing the law which serves to connect these with the actual, or so-called "direct" capacities, we can calculate these direct capacities. It is, however, very easy to modify slightly the connection of the bridge so that the direct capacity is at once given by the bridge setting.

There is furnished with this special bridge a tube socket, into which the triode fits, which is fitted with three prongs spaced on an equilateral triangle so that any pair of the three can be inserted into the terminals labeled C_x in Fig. 65. These three prongs connect to the filament, grid, and plate of the triode. Hence by inserting any pair into the bridge terminals, the total capacity between that pair of prongs is measured. This, as just pointed out, is considerably greater than the direct capacity which we are endeavoring to measure. The third prong of the special socket is left free, disconnected from the bridge, when this total capacity measurement is being made.

Now by putting in an extra connection on this special capacity bridge, whereby the unused prong is connected to the junction of the resistance arms of the bridge, the value of capacity given by the bridge is the actual, direct capacity. This extra connection is shown in Fig. 67. The two terminals a and b are those we have already assumed on the bridge. They connect to the two electrodes of the triode, the capacity between which it is desired to measure. Now it will be noticed that in Fig. 67 an extra terminal, C, has been fitted to the bridge, so positioned with respect to the other two that the prong of the special socket, which formerly was free, is now connected to the junction of the resistance arms R_1 and R_2.

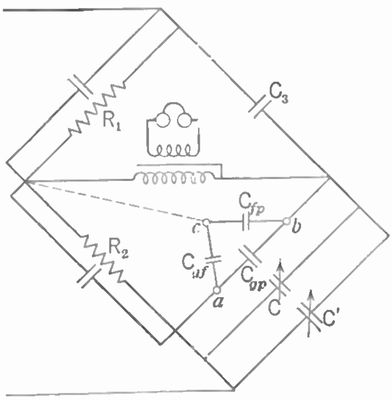


Fig. 67.

We have shown the special socket in position to measure C_{gp}, the two terminals a and b, of the bridge connecting to the grid and plate of the triode. Then the filament prong connects to the junction of R_1 and R_2.

Now it will be noticed that with this connection the condenser C_{fp} connects directly across the phones, and so when the bridge is balanced, there being no potential difference across the phones, the condenser C_{fp} is not excited, i.e., it draws no current and hence affects the balance of the bridge not at all. The condenser C_{gf} is connected in parallel with the resistance arm R_2; its effect on the bridge balance can then be compensated by an increase in part C_1 of the " loss condenser " and a corresponding decrease in part C_2. The amount by which the standard variable condenser, C, has to be reduced, to balance the bridge after the triode has been connected to the bridge, is the direct capacity between the two electrodes connected to terminals *a* and *b*.

It will be noticed, when the instruction for compensating condenser C_{gf} by an adjustment of the loss condenser C_1C_2, is analyzed, that this compound condenser is nothing but a scheme for changing the power factor of the 1 and 2 arms of the bridge, to make them correspond to the power factor of the 3 and 4 arms. When the bridge has been balanced, with the special socket in the bridge terminals, but no triode in the socket, not only are the capacities in the 3 and 4 arms proportional to the ratio R_1/R_2 but the power factor ratio of the 1 and 2 arms is the same as that of the 3 and 4 arms. Now connecting C_{gf} across the resistance R_2 of course upsets this power factor condition, so the relative power factors of the 1 and 2 arms must be brought back to the previous proportion if the bridge is to be again balanced and the restoration of power factor relation is accomplished by reducing C_2 and increasing C_1.

By means of this special capacity bridge, measure the various interelectrode capacities of the various types of triodes at present in use.

Measure the apparent capacities (extra prong not connected to the bridge) and the direct capacity. From the measured values of direct capacities calculate what the apparent capacities should be and compare with the values obtained from the bridge balance.

EXPERIMENT 13

Object.—Measurement of the electrical characteristics of a telephone receiver with and without diafram, with diafram clamped and with it free. Electrical characteristics of a loud speaker with and without diafram, with and without magnetic field (if the type permits). Effect of horn.

Analysis.—Both the head telephone and loud speaker are special forms of electric motors; their function is to make a diafram move back and forth, thus imparting compression and rarefaction to the adjacent air, in conformity with the shape of current supplied to the winding. How well they carry out their task can be judged to some extent by the measured values of resistance and inductance, these measurements to be carried out for various frequencies and various adjustments of the device.

We will first consider the telephone receiver, a simple cross-section of which is shown in Fig. 68. A U-shaped magnet, permanently magnetized in some way or other, has its poles N and S of soft iron so that the flux through them may be increased and decreased when current flows through the winding surrounding the pole pieces. A soft steel diafram D is clamped to a ring RR so that even though it bends under the pull of the permanent magnetic field from N to S, it does not distort sufficiently to touch the pole pieces.

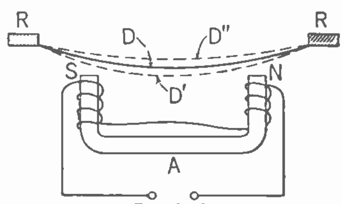


FIG. 68.

When current flows through the winding in such a direction that the magnetic field of the poles is strengthened, the diafram is further distorted to position D', and when the current flows through the windings in the reversed direction, thus weakening the magnetic field and decreasing the pull, the elastic quality of the distorted diafram pulls it back to some such position as D''. When alternating current flows through the winding the diafram must, therefore, execute a to-and-fro motion, and if this oscillating motion occurs with say fifty or more excursions per second, an audible sound is set up in the air adjacent to the moving diafram.

It must not be thought that the motion of an actual diafram has such an amplitude as suggested in Fig. 68. The amplitude is measured only in thousandths of an inch even when the phone is giving off a loud signal; for weak signal the excursion is probably measured only in millionths of an inch.

In the case of a loud speaker, giving off a very loud 100-cycle note (sufficient volume to fill a hall) the excursion is only a few hundredths of an inch, and for a 1000-cycle note of the same loudness the amplitude is measured only in thousandths of an inch. The pressure on the air increases and decreases from its normal value (14.7 lb. per sq. in.) by possibly one ten thousandth of a pound per square inch. For sounds of the intensity of ordinary speech, the change in air pressure, due to the sound wave, is only about 0.000001 lb. per sq. in.

The c.c. resistance (i.e., wire resistance) of an ordinary radio receiver head phone is about 1000 ohms. The a.c. resistance will of course be more than this, owing to the losses in the iron of the pole pieces and in the diafram. These losses increase with the frequency (hysteresis with the first power and eddy current with the square of the frequency) so that we may expect the a.c. resistance to increase quite rapidly as the frequency is increased through the audible range.

In order to find what part the diafram plays in the action of the magnetic circuit, the diafram may be removed and the effective resistance and inductance of the electromagnet may be measured. It will not be necessary to take many readings through the frequency range because the two factors change along smooth curves, when plotted against frequency. Some telephone magnets have solid pole pieces and some are laminated; the former will show greater changes in both R and L than the latter, R increasing and L decreasing as the frequency is raised.

If now R and L are measured through the same frequency range as before, with the diafram in place but clamped so that it cannot move, the resistance and inductance should both change with frequency at a more rapid rate than with no diafram. The reduction in the air gap of the magnetic circuit (the diafram reduces the air gap greatly) results in a greater flux being set up by the alternating current in the windings, hence a larger resistance and inductance. Offsetting the increase in inductance is the effect of the eddy currents in the diafram. At high frequencies the diafram eddy currents act as a magnetic shield to keep the alternating flux out of the diafram, thus making the inductance nearly the same as it is without the diafram.

If now the diafram is unclamped, so that it can move under the action of the alternating current in the winding, another action takes place. The diafram, moving back and forth synchronously with the increase and decrease in magnetic field strength, generates different hysteresis and eddy current losses than it did when it was stationary; whether they are greater or smaller than in the former case depends upon the relative phases of the motion of the diafram and the change in field strength.

It might be thought that the motion of the diafram would be exactly in phase with the strength of magnetic field; i.e., strongest field result in greatest velocity, etc., but such is not the fact. The diafram is a mechanical system which has mass, and an elasticity which tends to hold the diafram in a certain position. Such a system always has a frequency of motion at which a given force results in greatest amplitude of motion. This is called the resonant frequency of the system. The sharpness of resonance of the diafram's motion is affected by the losses incurred by its motion, such as frictional losses in the clamping ring around the rim of the diafram, viscosity losses in the material of the diafram itself, and energy imparted to the surrounding air.

The ordinary telephone diafram has a natural frequency at about 800 vibrations per second. Below this frequency, the motion (velocity) of the diafram leads the field strength by about 90°; that is, when the field is strongest the diafram is pulled down as close to the pole pieces as it can get. But at the extremity of amplitude any body executing harmonic motion is at rest; its maximum velocity, towards the pole pieces, occurred one-quarter of a cycle earlier.

If the frequency of current through the windings is the same as the natural frequency of the diafram, the amplitude is a maximum and the motion of the diafram is in phase with the magnetic field. As the frequency of the current in the winding is raised higher than the natural frequency, the amplitude of the diafram rapidly diminishes and the motion of the diafram lags 90° behind the magnetic field.

This peculiar change in the relative phase of diafram motion and magnetic field strength might be expected to be reflected into the electrical characteristics of the winding, as measured in the bridge, and such proves to be the fact. In Fig. 69 are shown in dashed lines the inductance and resistance of the phone with the diafram clamped, and in solid lines the same characteristics with the diafram free to vibrate.

A peculiar condition occurs for a frequency somewhat above the resonant value; *the effective alternating current resistance is less when the diafram is moving, giving off energy in the form of sound, than when it is clamped so that it can give off no sound energy.* It would seem that the motion of the diafram actually supplied energy to the magnet winding, but of course such cannot be the fact. The explanation of this curve brings out a seldom emphasized fact, the extremely low efficiency of the telephone receiver as a sound generator. Although the moving diafram is actually giving off energy in the form of sound, the extra resistance (positive) *which must be introduced into the windings by the required energy transfer from the electrical circuit* is much less than the decrease in hysteresis and eddy current losses in the moving diafram, compared to these losses when the diafram is clamped tight.

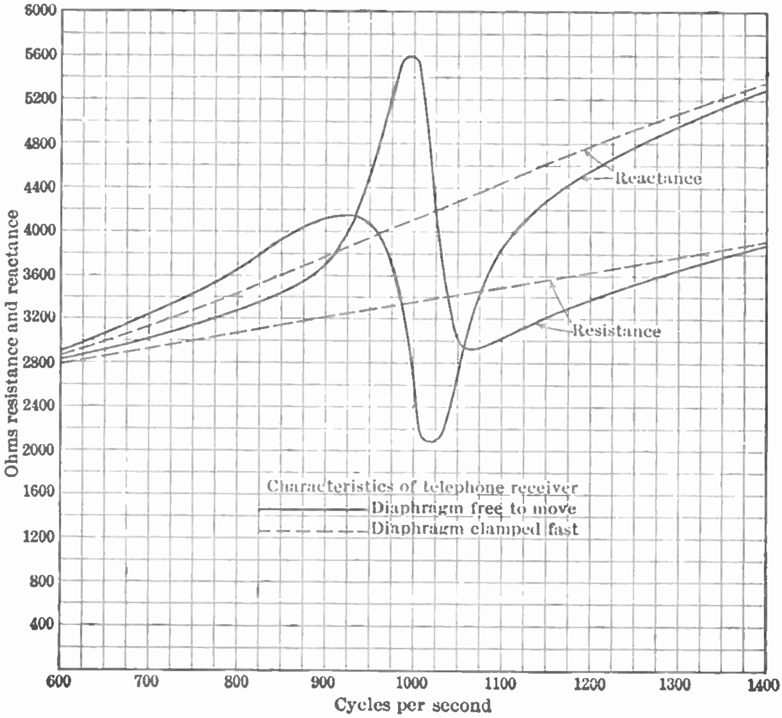


FIG. 69.

The measurement of the sound-generating efficiency of a telephone is a difficult one, so there are not many published data available, but it seems likely that the ordinary telephone receiver does not transform into sound more than 1 per cent of the electrical power supplied to its windings, and at frequencies far removed from the resonant frequency of the diafram even this figure is much too high.

Many of the loud speakers in use employ a moving diafram scheme similar to the head phone just analzyed, but add a properly designed horn to improve its efficiency as a sound generator. The reason for the extremely low efficiency of the vibrating diafram as a sound generator is a result of the air " running away " from the diafram when this moves forward to form the compression wave, and of air " running in " when the diafram, moving backward, is endeavoring to set up a rarefaction wave. Now if the air in front of the diafram is properly confined, the diafram's motion is much more effective in compressing and rarefying

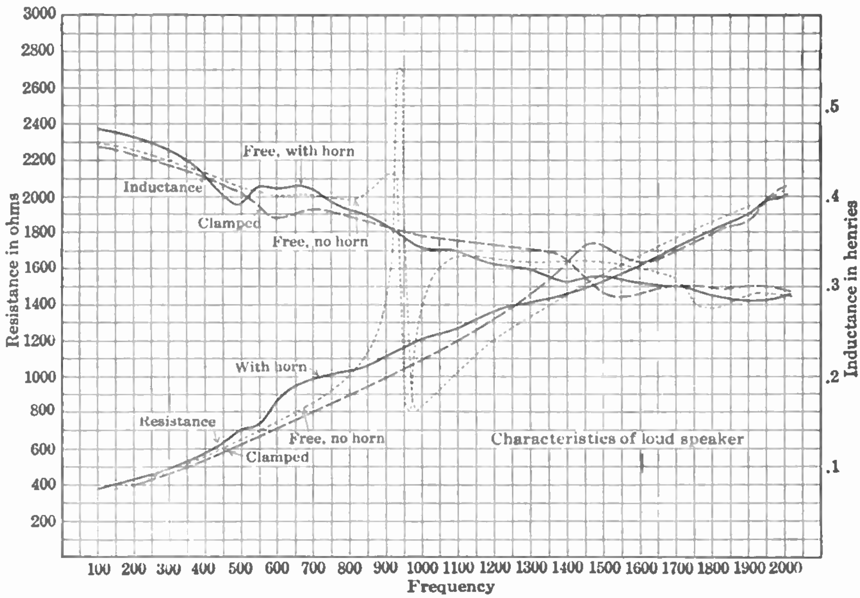


Fig. 70.

the air, and so a given motion sends off much more sound energy. Of course such a horn generally brings into the sound response some of its own frequency preferences so that the quality of speech or music may be considerably distorted by its action.

The effect of such a horn on the sound-producing qualities of the system is shown to some extent by the electrical characteristics measured in the bridge. In Fig. 70 are shown the measured resistance and inductance of such a loud speaker with the diafram clamped, diafram free with no horn, and diafram free with horn. With no horn the excessive motion of the diafram at the resonant frequency shows a great increase in resistance, and incidentally a very intense sound is given off by the speaker under this condition. It is an efficient sound generator (comparatively) for this one frequency.

When the horn is fitted to the speaker, the resonant response entirely disappears, but the effective resistance has been appreciably raised, especially for the lower frequencies. It is at these frequencies that this speaker without a horn gives almost no sound response whatever.

The more recent " moving coil " or " dynamic " speaker is much more efficient than the average moving diafram type. The moving coil itself has only about 10 ohms resistance (generally) so that it requires a transformer to make it take power efficiently from the output tube of the radio set. This transformer should generally then be treated as a part of the loud speaker.

Just as the moving diafram type requires that the permanent magnet supply its poles with a field much stronger than that set up by the windings (to prevent frequency doubling) the dynamic speaker coil requires a strong magnetic field in which to move. It would move, of course, with no current at all in the c.c. exciting winding, but the sound response would be very weak and would have a frequency twice that of the current flowing through the coil. The eddy current which the moving coil sets up in the faces of the magnet, and to some extent the hysteresis loss, will exist whether the magnet is excited or not, but the force produced in the moving coil will be very small, so no appreciable sound will be emitted. The change which occurs in these losses when the magnetic field is excited can be measured by getting the R and L of this coil with the exciting winding first not excited and then with the winding excited, in both cases having the coil clamped to prevent motion.

The coil may then be freed and its characteristics be obtained, with no paper diafram attached, and finally the moving coil may be measured when functioning normally.

It will be found that the increase in resistance due to the emission of sound by the speaker is much greater with this type of movement than with the moving diafram. This is particularly true at the lower frequencies, and is undoubtedly due to the larger amplitudes of motion available to the moving coil, compared to the possible amplitude of the ordinary diafram.

Using a bridge similar to that used in Exp. 10, measure the characteristics of a telephone receiver, for the various conditions analyzed, and also those of at least one type of loud speaker. The curves with moving diaframs are never smooth but have all sorts of humps due to partial resonance at various frequencies. If accurate results are desired, it is necessary to get measured points at a great many frequencies.

At first the characteristics should be measured with various currents, at a fixed frequency. If the inductance and resistance show appreciable changes with current variation, the current through the phone must be held constant when making the several frequency variation runs.

EXPERIMENT 14

Object.—Calibration of hot wire ammeters and thermocouples. Variation of resistance with current. Efficiency of conversion in the thermocouple.

Analysis.—Practically all radio measurements (with the exception of frequency) are made by means of ammeters and resistances; a radio frequency wattmeter is a practical impossibility, and a radio frequency voltmeter is almost never used because it requires too much power. (This statement will be qualified in the discussion of Exp. 22.) In view of this fact, it is important to have methods of checking the accuracy of the ammeters and resistances used in the radio laboratory.

As the force acting on an alternating current ammeter must be in the same direction, even when the current is reversed, it follows that an ammeter suitable for radio frequency measurements must be one in which the force varies with the square of the current. The iron vane ammeters, the induction type ammeters, the hot wire meter, the dynamometer type of meter, and the thermocouple meter all obey this law, and so are suitable for alternating current measurements. But at the high frequencies used in the radio field many of these types have serious defects, so that actually only the hot wire and thermocouple types have found appreciable application.

In the hot wire type, a thin wire is heated by the current to be measured and of course expands when heated. This expansion serves to actuate the finger, which moves over a scale properly calibrated in amperes. The expansion of the wire is of course too small to result in much motion of a finger; some kind of mechanical multiplying device must be used to make the finger move, say, over a 5-inch scale when the hot wire itself has expanded perhaps only 0.01 inch.

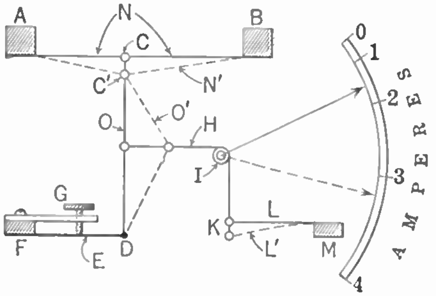


Fig. 71.

The general idea of the ordinary hot wire ammeter may be obtained from Fig. 71. The terminal posts A and B are mounted upon a base block; if this is of metal they are properly insulated therefrom. They are connected by a thin wire, *N*, about 3 inches long, through which, the current to be measured flows. Threaded on this wire at its mid point is a glass bead, *C*, to which is fastened a silk thread *O*. The other end of *O* is fastened to a stiff spring *D*, which is anchored on block *F*. This spring *D* can be adjusted by the screw *G*, the head of which is usually accessible from outside the meter case. Fastened to the middle of thread *O* is another silk thread *H*, which after passing around the shaft *I* of the meter once or twice, is fastened to a slender spring *L*, which serves to keep the two threads stretched taut.

When the wire *N* is heated it expands, and the threads move to the positions shown by the dotted lines. The motion of the finger is nearly proportional to the expansion of wire *N*, so that the scale of the meter, in current, is of the parabolic type, very crowded at the lower end of the scale. The expansion of the wire *N* is four times as much for 2 amperes as it is for one, so the finger moves practically four times as far on the scale for 2 amperes as it does for 1.

The zero setting of such a meter must be readily accessible, as the position of the pointer will evidently change with temperature, etc. If blocks *A* and *B* are mounted on metal which has the same temperature expansion coefficient as has the wire *N*, the effect is minimized, but generally this is not the fact. The set screw *G* evidently permits the ready adjustment of the zero setting.

The a.c. ampere is defined as that current which will heat a given resistance to the same temperature as will a c.c. ampere, so it follows that this meter can be calibrated on a c.c. circuit. The question arises, however, to one studying the peculiarities of very high-frequency currents, as to whether the resistance of the wire is the same for the radio frequencies, at which the meter is to be used, as it is for a continuous current. The answer to the question is, no. The resistance may be very different at one megacycle than it is for continuous current, owing to the skin effect existing at the higher frequencies.

In a c.c. circuit, the current distributes itself uniformly throughout the cross section of the wire, but at very high frequencies the current density at the center of the wire may be only a small fraction of what it is at the outside of the wire. The difference in current distribution results in a difference in resistance, the uniform distribution giving the lowest resistance. The factor by which the c.c. resistance must be measured to give the a.c. resistance increases with frequency and with size of wire, and decreases as the specific resistance of the material used for the wire increases.

Hence if we want a wire to have the same resistance at radio frequencies as it has for continuous current, it must be of some high resistance material, and must be of small diameter. In the table below is given the largest diameter of wire, of different material, which can be used, at a specified frequency, without having the a.c. resistance exceed the c.c. resistance by more than 1 per cent.

WIRE DIAMETERS

Largest wire (straight) which can be used without the high-frequency resistance exceeding the c.c. resistance by more than 1 per cent

Wave length, in meters	Diameters Given in Millimeters			
	Advance	Manganin	Platinum	Copper
100	0.30	0.29	0.13	0.006
200	0.46	0.40	0.20	0.045
300	0.57	0.50	0.27	0.09
400	0.66	0.60	0.30	0.10
600	0.83	0.75	0.37	0.15
800	0.98	0.88	0.42	0.20
1000	1.10	0.99	0.50	0.21
1200	1.20	1.10	0.57	0.22
1500	1.30	1.21	0.63	0.26
2000	1.52	1.38	0.73	0.30
3000	1.82	1.62	0.80	0.33

$$\text{Frequency} = 3 \times 10^8 \div \text{wave length.}$$

Thus if we want to use a piece of manganin wire for a hot wire ammeter good to 3×10^6 cycles per second (100 meters) and the resistance for the radio frequency current is not to be greater than 1.01 times the c.c. resistance, its diameter must not exceed 0.29 mm. (No. 30 wire). And if copper wire is to be used its diameter must not be greater than 0.006 mm., which is smaller than a No. 40 wire.

But if the wire must be so small, and of high resistivity, it evidently cannot carry much current without overheating. To construct an ammeter good for more than a very few amperes, therefore, it is necessary to use several wires in parallel. When this is done it will be found that the current does not divide evenly between the different wires, unless special care is exercised in their arrangement. They must be in some symmetrical arrangement with respect to a central axis; in Fig. 72 is shown the construction of a meter good for 50 amperes. Instead of using wires, very thin strips of a high-resistance alloy are arranged much like the staves of a barrel, and the heavy copper castings which serve as the barrel heads connect to the heavy terminal posts of the meter. The expansion of one of these strips is used to actuate the finger of the meter. The various parts of the meter are mounted on a marble slab.

As we attempt to use the hot wire idea for the construction of meters to measure weak currents, we soon reach a lower current limit beyond which it is impossible to go. For a given resistance the heat developed

FIG. 72.

varies as the square of the current, thus a 0.1-ampere meter with a given resistance would generate only 1 per cent as much heat as a 1-ampere meter. To offset this effect the lower range meter is made of smaller wire having higher resistance. However, even if we use a wire so fine as to be scarcely visible to the eye, it requires about 0.05 ampere to heat it sufficiently to move the pointer over a very small scale. The resistance of this meter would be about 10 ohms.

By using the thermoelectric voltage set up on a bi-metallic junction, however, we can build meters, good for radio frequency measurements, which will operate with much less current than the value given above as the limit for the hot wire ammeter. In this type of meter a wire is heated by the current to be measured, and this heated wire is in contact with a thermal junction. The thermal junction, heated, generates a continuous voltage which serves to operate a sensitive c.c. galvanometer. To get as much temperature on the hot wire (and hence on the junction) as possible, with little power, it is necessary to eliminate the cooling effect of air in contact with the wire. To carry out this idea the thermocouple and heater wire are mounted inside a small glass bulb, from which the air is afterwards exhausted. The arrangement is as given in Fig. 73.

The heater wire, *c*, is straddled at its mid-point by the junction, *d*; comparatively heavy wires go through the "press" and serve to hold the heater wire and thermocouple in place.

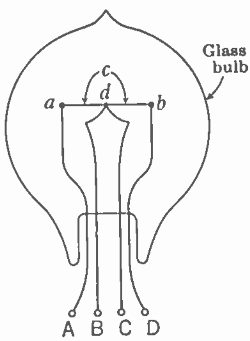


FIG. 73.

Without exceeding the safe temperature of the wire *c*, the junction, *d*, will generate about 8 millivolts; the temperature of the junction will be about 350° C, and of course the wire itself will be much hotter. It requires some thought to answer the very simple question as to how long the wire, *c*, should be. If it is made too short it will be cooled too much by the heavy wires at its ends, and if it is made too long its resistance is unnecessarily high. The joints *a* and *b* serve to cool the heater at its ends, and the couple itself, *d*, serves to keep it cool at its middle, so we may expect it to be hottest half way from the couple to either end. Here the temperature is about 500° C when the couple is at 350° C.

As these devices are offered on the market today the couple itself is the same whatever type of heater wire is used. The resistance of the couple from *B* to *C* (Fig. 73) is very close to 10 ohms, whereas the heater wires may vary in resistance from a small fraction of one ohm to a thousand or more ohms. Of course this thermocouple is of no use itself as a current indicator; it is merely a conversion device to change a.c. power into c.c. power. The c.c. meter used to operate from the power generated in the couple should have the same resistance as the couple itself, if the couple is to deliver a maximum power. The proper meter to use, therefore, is a millivoltmeter of about 10 ohms resistance. If a portable meter of the pivot type is used a full scale reading of about 2.5 millivolts is about as sensitive as can be obtained; if a semi-portable meter with suspension construction is used, about 0.25 millivolt full scale can be obtained. If a telescope and mirror instrument, of 10 ohms resistance, is used, a deflection of 10 c.m. (about the size of the scale of the meters just discussed) can be obtained with 0.05 millivolt.

Of course, when any of these meters is indicating a certain number of millivolts, the temperature of the couple is great enough to generate just twice the amount of voltage; one-half the voltage it generates is used up as internal resistance drop.

The resistance of the heater wire depends upon the current it is designed to carry; on page 9 of the text is given a table showing approximately the resistances and current ranges as couples are made today.

Now a very important point in connection with the use of hot wire ammeters and thermocouples is the constancy of their resistance. Both types of meters depend for their operation upon increase of temperature of their current-carrying wires, and generally the resistance of a wire varies with temperature. It is generally assumed that the resistance of the meter itself is independent of the current through it, but such is not actually the fact. In many of the measurements made in the radio laboratory, a certain current is obtained in the test circuit; resistance is then added in the circuit to decrease the current to a certain fraction of its original reading, and from the required value of this resistance, conclusions are drawn as to the constants of the circuit. These conclusions will be wrong if the ammeter itself appreciably changed its resistance when the current in the circuit was decreased. The heater wires in very low current thermocouples are generally of carbon, and for the higher current capacities are some sort of metallic alloy. We might conclude, therefore, that the low current heaters would decrease in resistance with increasing current and that the others would show in increasing resistance.

Typical results from commercial thermocouples are given here.

Couple No. T 376		Couple No. 227	
Current, amperes	Resistance, ohms	Current, amperes	Resistance, ohms
0.000388	693	0.108	0.300
0.000621	689	0.173	0.302
0.000831	684	0.254	0.305
0.000973	679	0.330	0.309
0.001122	675	0.410	0.314
0.00135	668	0.489	0.319
0.00148	664	0.557	0.327
0.00163	650	0.657	0.338

From these values it is seen that if accurate results are to be obtained in certain radio tests, allowance must be made for a considerable change in ammeter resistance, sometimes an increase and sometimes a decrease.

Using the arrangement of apparatus shown in Fig. 74, calibrate two hot wire ammeters and two thermocouples. A slide wire bridge is supplied with power from a storage battery of one or two cells connected to a potentiometer having a current-carrying capacity sufficient for the full scale of the larger of the two hot wire ammeters. A reversing switch permits rapid changes for reversed current reading. The voltmeter V, of about 3-volt range, is preferably of the center zero scale type. The resistance R should be of about the same resistance as the meter, A, being tested. It should carry safely the maximum current to be sent through the meter A. G is an ordinary bridge balancing galvanometer.

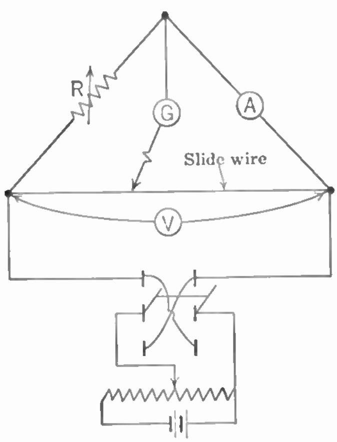


Fig. 74.

For each meter get about eight readings of voltages V and bridge setting, taking readings for "direct" and "reversed" power supply. R must be accurately known. If not, it must be measured. The value of V for which readings are taken should be properly selected to give reasonably well-separated points on the scale of the meter being tested.

Note whether the zero reading of the meters changes as current is passed through them, and if the zero does change find some experimental way of allowing for the change so that the ammeter readings may be taken as accurate (allowing for the calibration which the test gives).

From the bridge setting the resistance of the meter is given. From this (in combination with the known value of R) and the reading of V, the current through the meter is calculated. The bridge connection resistances must be so low as to be negligible compared to the meter resistance, in this scheme of calibration.

Plot calibration curves and resistance curves for the four meters tested.

For the two thermocouples calculate the efficiency of conversion, at about four points on the scale, equally spaced.

EXPERIMENT 15

Object.—Study of rectifiers. Copper oxide, crystal contacts, gas tubes, with and without heated electrodes. (Diodes taken up later.)

Analysis.—The rectifier, or one-way conductor, serves many purposes in radio apparatus, and according to the service for which it is intended may be of high resistance and low current capacity or the reverse.

It is seldom that the rectification is complete, with any of the commerical forms of rectifiers; they simply carry current more readily in one direction than in the other, that is, they show assymmetrical conductivity. Now practically any contact surface, between different conducting materials, has this property to a greater or less extent; thus the contact between a hot and cold surface of even the same material shows some rectification.

Of course, to be of much value the contact must show a much greater conductivity in one direction than in the other, say ten times as much. Even when such a contact is found, it is generally useful in a very limited range of voltage. Thus a given contact might rectify in one direction for voltages (a.c.) less than 2 volts, not rectify at all at 3 volts, and rectify in the opposite direction for higher voltages.

The exact mechanism by which contacts rectify has been the subject of much speculation; it seems reasonable, in view of the theories of modern physics, to assume that it is merely a question of the ease with which electrons are able to escape through the surface of the contacting materials. The amount of energy required to carry an electron through the surface of a conductor (sometimes called the *surface work*) is different for different materials, and even for different surfaces of the same material if this is crystalline in its structure, so that when a voltage is impressed across the contacting surface it seems likely that the electrons will flow across more readily in one direction than in the other.

The high-frequency currents which are set up in a radio receiving circuit are incapable of giving an audible signal in the phones until they have been passed through a rectifier. The intensity of voltage available, to impress on the rectifier, in this case is generally a fraction of 1 volt; it may possibly be a few volts but such a case is unusual. The amount of rectified current required for actuating a head phone is only a few microamperes, so that this service requires a rectifier which will give a few microamperes of current when a signal measured in fractions of a volt is impressed.

There are many crystal contacts which serve well for this purpose; sometimes two crystals in contact with each other are used and sometimes a fine metallic point (like the point of a needle) is lightly pressed against the crystal surface. In Fig. 75 are shown the current-voltage curves obtained from a double crystal contact, zincite pressing lightly against a crystal of chalcopyrite. These curves are obtained by impressing various voltages across the crystal contact, first in one direction and then in the other, and reading the resultant current. The solid line curve is for a good rectifying point and the dashed curve is for a point which rectifies not so well. These curves have been taken for voltages greater than generally available in radio receiving circuits; the important part of the curve where it should show good rectification is the part between plus and minus 1 volt.

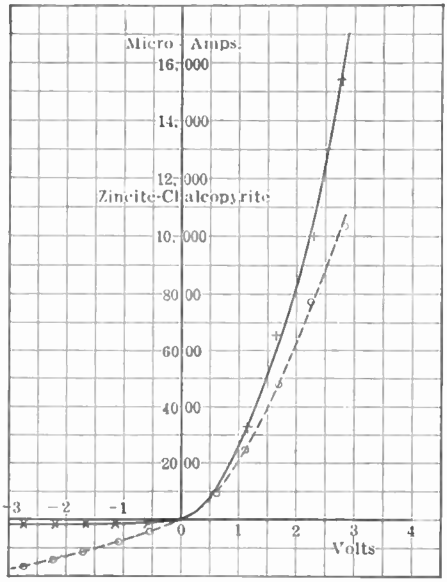


Fig. 75.

In Fig. 76 is shown the rectification between a steel needle point and a silicon carbide crystal; here again the characteristics of a good and poor point are shown. This type of contact gives its best rectification if a constant polarizing voltage (in this case of about 1.2 volts) is impressed. The rectification is by no means as good as for the crystal of Fig. 75, but the silicon carbide rectifier has the advantage of being less easily spoiled by strong signals, jarring, etc.

In Fig. 77 are shown the a.c. and c.c characteristics of a crystal contact for strong and for weak signals. The a.c. characteristic is obtained by impressing a known alternating voltage across the crystal and reading the current which flows with a c.c. galvanometer. This of course will read only the amount of rectified current which flows through the contact.

The points on the a.c. characteristic can be calculated for any point by assuming any value of a.c. voltage; from the c.c. characteristic the current for any instantaneous value of this voltage is obtained, and from the curve so obtained the average value of current is measured, by planimeter or similar scheme. This idea is illustrated in Fig. 78.

Another type of rectifier which is very useful is represented by the copper-copper oxide contact. A plate of copper, properly cleaned, is oxidized by a definite process. A plate of some soft metal is then pressed tightly against these oxidized surfaces and the rectified current flows between these plates and a lug of the copper plate which has been cleaned of oxide. The rectifying properties of such a surface are shown in Fig. 79; it will be noticed that the scale for reversed current is in milliamperes, whereas the direct current is in amperes. The rectification is thus nearly perfect in the voltage range shown.

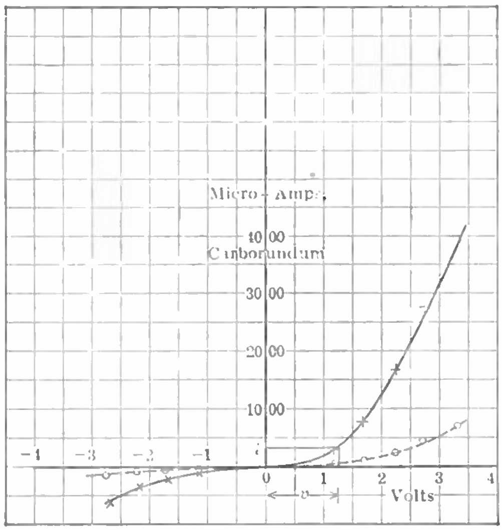


FIG. 76.

In common with all types of rectifiers the copper-copper oxide is spoiled if too high a voltage is impressed. It heats up too much and the oxide cracks away from the copper, thus ruining the contact. This rectifier, it will be seen, is serviceable where currents of the order of an ampere are required from an a.c. supply of a few volts. If a higher voltage source is to be rectified, the proper number of these units must be connected in series.

Where a current measured in milliamperes is desired, from a circuit of from 1 to 300 volts, a tube containing rarefied gas has been much used. In this type of tube (generally marketed under the trade mark "Raytheon") the rectifying properties of a rarefied inert gas in a bulb having differently shaped electrodes is utilized. By suitably placing the electrodes with respect to one another, and suitably forming these electrodes, it is possible to build a tube which will carry a considerable fraction of an ampere (say 0.2 ampere) with only about 150 volts across the tube, and will pass practically no current in the reversed direction until perhaps 300 volts is impressed. In Fig. 80 are shown conventionally the current-rectifying properties of such a tube.

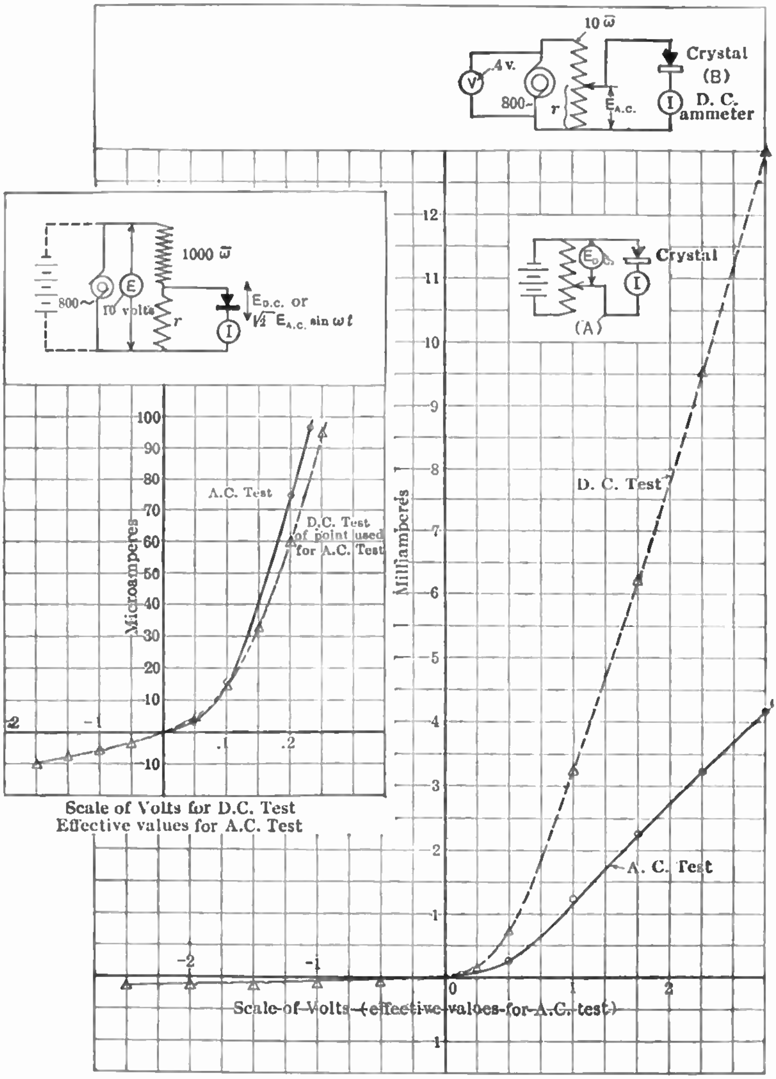


FIG. 77.

All tubes containing appreciable amounts of gas are more or less unstable; when conduction occurs, owing to progressive ionization of the gas at voltages above that required for breakdown, the tube is likely

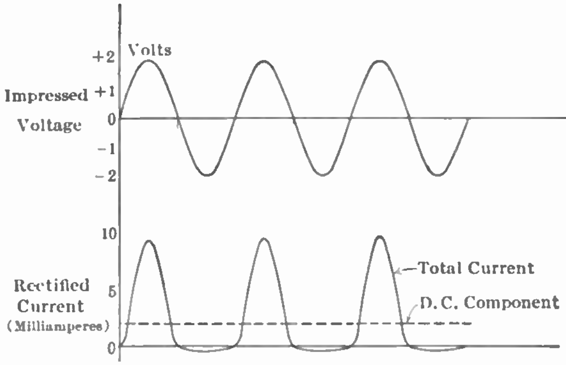


FIG. 78.

to short-circuit the power supply unless a proper amount of stabilizing resistance or reactance is in series with the tube to limit the current flow.

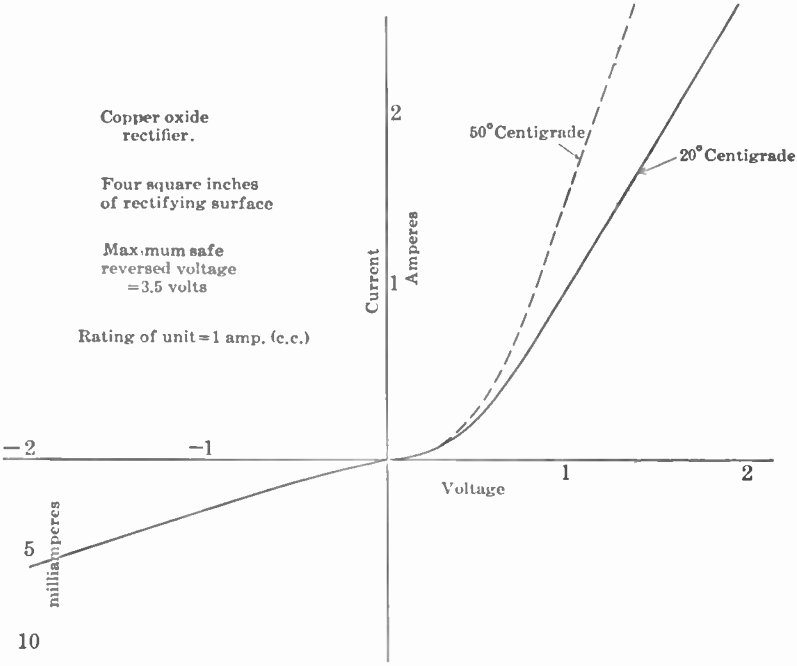


FIG. 79.

For satisfactory operation this type of rectifier requires the proper amount of rarefied gas present; if too much voltage is used on the tube it will overheat, the gas pressure will be changed, and the tube possibly spoiled.

The type of tube just analyzed requires quite a high voltage between its electrodes, it will be noticed, before it carries appreciable current, even in the conductive direction. Another type of gas rectifier which overcomes this difficulty uses for one electrode a heated filament, from which electrons are evaporating. These evaporated electrons are sufficient to start the ionization of the gas in the tube at very low voltages, when the proper polarity of voltage is impressed between the hot filament and the other electrode. This polarity requires the hot filament to be negative with respect to the other electrode. When the polarity of the impressed voltage is reversed no current flows because there are available no electrons for ionization. Those evaporating from the hot filament are forced back into the filament by the impressed voltage, and the other electrode, cold, is not evaporating any electrons.

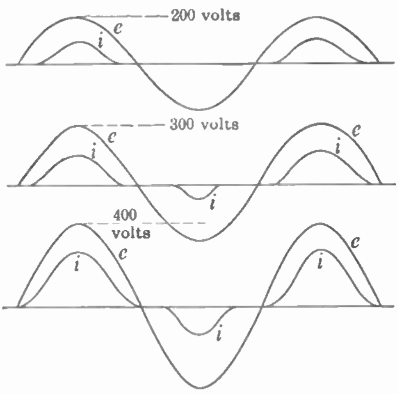


FIG. 80.

In tubes of this nature it is quite possible that the hot filament may be spoiled, in its electron-emitting qualities, by the action of the positively charged gas ions which continually bombard it during that part of the cycle when the tube is carrying current. The type of filament, kind and pressure of gas, and voltage condition in the circuit all have to be properly considered if the filament is not to be ruined by this bombardment in a very short time. Generally neon or argon gas are used in these tubes (some of which are marketed under the trade name "Tungar") at a pressure of about one-half an atmosphere.

The mercury arc rectifier (of Cooper Hewitt) uses this idea, but the heat required to emit the electrons from the negative electrode (a pool of mercury) is supplied by the rectified current flowing through the tube. A white hot spot on the pool of mercury occurs where it is being bombarded by the positive ions, and it is from this hot spot that the electrons which produce the ionization of the mercury vapor originate.

In the latest type of rectifier of this general type, a hot filament is used for one electrode, and mercury vapor at low pressure is the gas which when ionized helps to carry the current. A grid is placed between the plate (cold electrode), and the filament and this grid serves to control the start of the ionization of the mercury vapor. Of course, in this tube as in the others, the ionization takes place only when the hot filament is negative, but whereas in the previously discussed tubes the current flow started when the plate reached a certain positive voltage, in this grid-controlled rectifier the potential of the grid serves as the dominating control. By varying the grid potential, if this is from a c.c. source (or by varying either its amount or phase if a.c.), the amount of rectified current which the tube delivers to its load may be controlled.

The action of this grid is not the same as that of the grid of an ordinary triode; after the grid potential has once permitted the starting of the ionization phenomenon, it is helpless to control the magnitude of current to the plate for the rest of that alternation of the power supply; during the alternation when the plate is negative the ionization of the mercury vapor ceases, so the grid regains its control for the beginning of the succeeding alternation. (Some of these special rectifiers are being sold under the trade name of " Thyratron.")

Get characteristic curves, both c.c. and a.c., for a crystal rectifier such as has been used in radio receiving circuits. Get curves for a good, and a poor, point. Note how easily a good point is lost, by jarring the crystal.

Get curves for a copper oxide rectifier. In this test it will be necessary to allow for temperature variation if consistent results are to be obtained. At the higher voltages, after having set the conditions for the desired reading, take off the power, let the plate cool down, and then get readings as soon as the switch is closed, or some other consistently uniform procedure may be followed. Note whether the plate heats.

Get curves, both a.c. and c.c., for a " Tungar " or similar bulb. In this case have a storage battery in the circuit, of voltage rating for which the outfit is designed.

Study the action of the " Thyratron " sufficiently to become familiar with its characteristics.

In getting the a.c. tests called for above, it is most essential that the a.c. voltage supply have sufficiently close regulation so that the wave form impressed on the rectifier retain its sinusoidal form during the complete cycle. Part of the cycle current is being drawn and part of the cycle the power supply is not delivering any current. This is likely to result in the voltage wave, impressed on the rectifier, being flat topped during the conductive part of the cycle. In Fig. 81 this idea is brought out; a low voltage is required, say for the copper oxide rectifier. A step-down transformer with a secondary current rating of several times that carried by the rectifier is advisable. To vary the voltage impressed on the rectifier a resistance might be used as in *a* or *b* of Fig. 81. Neither of these schemes will work at all; the amount of rectified current obtained will be much less than it should be, because during that part of the cycle when the rectifier is carrying current, the voltage across it will be much smaller than it should be. If the scheme of Fig. 81*c* is used, however, the a.c. test will check with the values predicted from the c.c. test results. The voltage on the rectifier will be sinusoidal; its value may be found by measuring the voltage impressed on the high side of the transformer and allowing for the known transformer ratio. The amount of voltage is varied, not by a resistance in either circuit of the transformer, but by varying the field current of the alternator.

In case an alternator is not available for the test, so that the laboratory power line must be used, it is possible to use the arrangement of

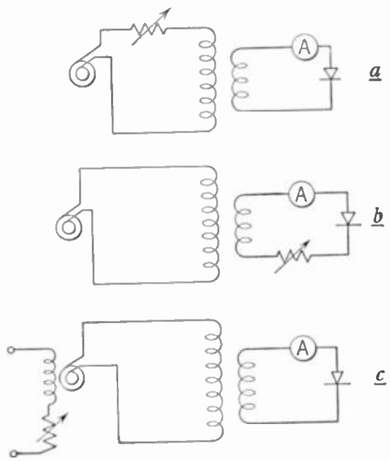


FIG. 81.

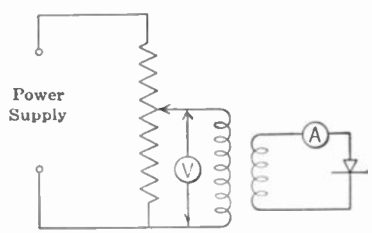


FIG. 82.

Fig. 82, but the potentiometer must be of low resistance. To test whether it is sufficiently low, put such a resistance on the secondary of the transformer (without the rectifier) that a secondary current flows of magnitude somewhat greater than the maximum current to be taken by the rectifier. If, on opening and closing the secondary circuit, the voltmeter across the primary shows no change, the potentiometer will be suitable.

For the crystal rectifier, and copper oxide rectifier, construct curves of the current form during an a.c. cycle and compare the average value of this with the value obtained in the a.c. test, for that voltage which corresponds to the voltage assumed in drawing the current form.

EXPERIMENT 16

Object.—Study of various types of wave meters. Use of a buzzer generator and wave meter for measuring self and mutual inductance and capacity.

Analysis.—The construction and performance of wave meters was taken up in the Introduction of this test, and the student should study the material there given before carrying out the work required in this experiment.

In this laboratory work, the student is to observe carefully the construction of several classes of wave meters, noting about what value of capacity is used, what type of condenser it is, the kind of scale (frequency or wave length) calibration the meter has, the method or methods of detecting resonance, etc. Notice whether the type of condenser and the kind of scale calibration seem to have any predictable relation.

At least four types of wave meters should be studied. One, similar to the Telefunken, which has a wave range from 100 to 6000 meters and various extra equipment which makes it a very useful laboratory meter; one designed for short waves, say from 5 to 10 meters; one for very long waves, say up to 20,000 meters, and some special one like the Kolster combined wave meter and decremeter.

These meters should be opened up, if necessary, so that their construction may be investigated. Great care must be exercised not to jar the meter or bend the condenser plates in the slightest amount, for of course this would entirely spoil the calibration of the meter.

Note whether the calibration of the meter is different when the different resonance indicators are used. If there is a difference, note whether the difference is constant throughout the range of the meter.

The laboratory standard wave meter should of course never be opened for inspection nor used in any way except for the calibration of the other laboratory wave meters.

The buzzer wave generator is a scheme for getting a few milliwatts of damped, high-frequency power from a low-voltage c.c. source. Whereas measurements made by the power from such a source do not have as much precision as those made by a power source giving continuous oscillations, its use is simpler and often more convenient.

The method of operation of the buzzer generator may be understood with the help of Fig. 83. A small U-shaped soft iron magnet is wound with several hundred turns of wire. A soft iron armature A is pivoted at B and some sort of adjustable spring action keeps it pressing against the contact screw D. The electrical circuit is through the winding, armature, and contact screw, in series. If current is passed through such an arrangement, it acts just like the ordinary electric doorbell; the armature vibrates at a frequency fixed by its mass and the tension on its spring adjustment.

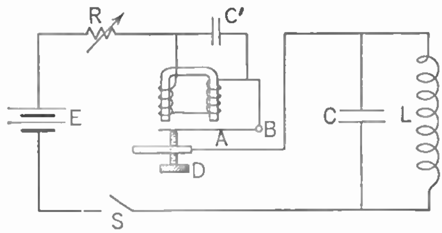


FIG. 83.

The contact surfaces of armature and screw D are necessarily covered with some non-oxidizable metal such as platinum, platinum-iridium alloy, or sometimes tungsten.

It will be noticed that the battery E, instead of being connected directly to the buzzer, is connected through an adjustable resistance R, switch S and coil L. When the switch S is closed, current starts to increase through the circuit on a logarithmic curve $i = (E/R)\left(1 - \epsilon^{-\frac{Rt}{L}}\right)$ in which R is the total resistance in the circuit and L is the total inductance. When the current reaches a certain value the armature is attracted sufficiently to the electromagnet to pull away from the contact screw D. If sparking does not occur at the contact (it will not if the buzzer is operating properly) the circuit made up of L and C is cut off from the buzzer and battery at this instant. The condenser C', shunted across the electromagnet, is for the suppression of such spark as the contact may tend to set up. It must not be too small, neither should it be too large, to get the best results as far as non-sparking is concerned. Its proper size is determined by the inductance of the electromagnet and the vibration frequency of the armature.

Frequently a resistance is used in parallel with the electromagnet instead of the condenser C'. Such a scheme generates somewhat more high-frequency power than if a condenser is used, but it may cause somewhat more sparking at the contacts.

To get a clear musical tone, free from a scratchy note, from such a buzzer, the right value of resistance R, the right separation between armature and pole pieces, and the right tension on the spring adjustment must be used. Further, the contact points must be kept smooth and clean.

Now returning to the cycle of events in the buzzer operation. At the time the contact opens, there is a certain current I flowing in the circuit, and of course a certain voltage V on the condenser C. This means that the battery E has stored energy in the L-C circuit, an amount equal to $LI^2/2$ in the coil and $CV^2/2$ in the condenser.

When the contact opens the current in the coil will tend to fall to zero, but in doing so it charges the condenser C to a greater voltage than it already has. We will calculate this voltage for a typical case. Suppose $L = 1$ millihenry, $I = 0.5$ ampere, $C = .001$ microfarad, and $V = 2$ volts. (This 2 volts of course must be somewhat less than the voltage of battery E.)

The energy $LI^2/2$ is equal to $\dfrac{.001 \times .5^2}{2}$ joules $= .000125$ joule of energy.

The energy $CV^2/2$ in the condenser is $\dfrac{10^{-9} \times 2^2}{2} = 2 \times 10^{-9}$ joules, which is negligible compared to the energy in the coil, hence we neglect it in further calculation.

When the coil energy has all passed over into the condenser, this must be charged to such a voltage that $CV^2/2 = .000125$. Calculation shows that the voltage is 500 volts. Now this voltage will tend to produce sparking at the contact points, and will, unless they are sufficiently separated at the time the voltage occurs. The frequency of oscillations in the LC circuit is given by $f = \dfrac{1}{2\pi\sqrt{LC}}$ = about 160,000 cycles per second. The high voltage across C occurs one-fourth of a cycle after the oscillations start, or only about 1/600,000 of a second after the contacts open. In this short time the gap cannot be very wide (between armature and contact screw) so sparking across the minute gap is likely to occur.

It looks at first as though a condenser might be put across the gap, to stop this sparking tendency. This, however, would leave the buzzer-battery-resistance R circuit electrically connected to the LC circuit (through the condenser across the gap), so that only very weak oscillations, of low frequency, would be set up.

If the spring tension is reduced, so that the armature will pull away from the contact screw with only 0.1 ampere in the electromagnet, the voltage set up across C will be somewhat less than 100 volts, so sparking is not so likely.

The oscillations set up in the LC circuit will die away at a rate depending upon the resistance in the circuit. The reactance of the coil will be 1000 ohms ($2\pi fL$) and if we assume a power factor of coil of 2 per cent and of condenser 1 per cent, the resistance of the oscillatory circuit is 30 ohms. The decrement which determines the rate of decay of oscillations is equal to $R/2fL$ = about 0.1. The oscillations will have died to 1 per cent of their original strength in a number of cycles given by $\frac{4.6 + \delta}{\delta} = 47$. And 47 oscillations (frequency 160,000 per second) will occur in 1/3000 of one second. Hence if the armature of the buzzer is vibrating 500 times a second, let us say closed 0.001 second and open 0.001 second, then the high-frequency oscillations have decayed to practically zero in only one-third of the time the contact is open. Hence when the contact closes again the oscillatory circuit is electrically dead, ready to repeat its previous performance.

The amount of high-frequency power generated by the buzzer is rather small. If we assume the buzzer adjusted to open on the 0.1 ampere of current, the energy in the coil is $\frac{0.001 \times 0.1^2}{2} = 0.000005$ joule, and this amount is changed into high-frequency energy 500 times per second. The amount of power is therefore 2.5 milliwatts.

If then, in some laboratory work, a wave meter is being excited from a buzzer generator, it must not be expected that the hot wire ammeter will show any deflection. It requires about 10 milliwatts *in the hot wire meter itself* to show a readable deflection, so it is quite evident that in experiments employing a buzzer wave generator for power supply the wave length must be measured in the wave meter by detector and phones, rather than by hot wire ammeter.

Having a buzzer, a calibrated wave meter, and a known inductance (or known condenser), any condenser or inductance suitable for use in radio circuits can be readily measured. The unknown condenser is connected to the known inductance and then arranged for excitation from the buzzer, that is, in Fig. 83, C is the unknown condenser, and L is a coil of known inductance. The wave meter is coupled loosely to this oscillatory circuit and its setting varied until resonance with the oscillating circuit is indicated by a maximum sound in the phones. (Hot wire ammeter cannot be used for resonance indicator.) Now for any setting of the wave meter its calibration will give either frequency or wave length, and to this value there corresponds a definite product, LC. This relation is obtained either from the equation

$$\lambda_{\text{meters}} = 1884\sqrt{LC} \quad \ldots \ldots \quad (38)$$

or

$$f_{\text{cycles per second}} = \frac{10^6}{2\pi\sqrt{LC}} \quad \ldots \ldots \quad (39)$$

in which formulas L and C are both given in micro-units (microhenrys and microfarads).

Now when the *LC* of the wave meter is known and this circuit is in resonance with the oscillatory circuit connected to the buzzer, the *LC* of this latter circuit is also known. Hence if either *L* or *C* of the circuit is known the other quantity can at once be calculated.

If then another coil is to be measured, this is now substituted for the known coil, the new *LC* again determined by wave meter, and as *C* is now known the new value of *L* can be calculated.

Various arrangements of the unknown *L* or *C* in combination with the known quantities enable the experimenter to measure quantities which he otherwise might not be able to. Thus suppose one of the condensers is so large that it in combination with the smallest known *L*, gives a *LC* product beyond the range of the wave meter. This large condenser can be connected in series with one of the known condensers and this combination connected to one of the known coils. The *LC* can then be measured on the wave meter and the value of *C* calculated. From the value of *C*, the known value of one of the condensers of the series combination, and the known law for capacity of condensers in series, the capacity of the large condenser can be calculated.

Short wires must be used in connecting the buzzer-excited circuit, because the connecting wires are in parallel with the condenser to be measured and so augment its capacity.

It will be found that a considerable signal will be heard in the wave meter phones wherever it is tuned. This is due to induction directly into the wave meter from the buzzer itself. This sends off pulses of electric and magnetic fields of the same frequency as the tone of the buzzer, and these pulses produce sound in the wave meter all over its scale, being generally loudest at the smallest setting of the wave meter condenser. This noise will be heard even if the wave meter crystal is not adjusted on a rectifying point, whereas the desired high-frequency signal (from the *LC* part of the buzzer set up) will be heard only if the crystal in series with the wave meter phones is rectifying properly.

The signal which the listener hears is about as indicated in Fig. 84. Curve *a* represents the interfering signal from the buzzer; and curve *b* represents the response of the wave meter to the desired high-frequency signals. The solid line is the signal the listener actually hears. By weakening the coupling (taking the wave meter farther away from the buzzer circuit) the total signal becomes weaker, curve *b* also becomes narrower, and when the best coupling is obtained the wave meter is silent except for two or three divisions of its scale. This response is the peak of the *b* curve and gives the frequency of the *LC* circuit which is the quantity desired.

To get the best ratio of desired signal to interference, the coil of the wave meter should be kept in the position which gives maximum coupling with the test coil, for a given separation. This is illustrated in Fig. 85. The wave meter is indicated by either A or B. Position A (wave meter coil and coil L coaxial) will give a better ratio of signal to interference than position B. By rotating the wave meter coil from position A to position B the signal is cut down more than the interfering signal.

Many wave meters are themselves equipped with a buzzer, so that the wave meter itself may be used as a calibrated buzzer generator. In this case the unknown condenser is connected in series with the known coil, and a crystal detector and phones, in series, are connected across the condenser. The setting of the wave meter (and hence the frequency of the wave it sends out) is changed until a maximum signal is heard in

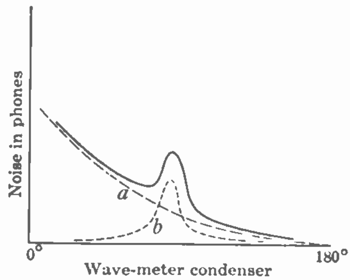


Fig. 84.

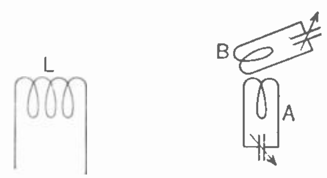


Fig. 85.

the phones. The LC of both circuits is then the same, and so the unknown quantity can be calculated.

Measure the inductance and capacity, by the method of this experiment, of coils and condensers which have previously been measured at audio frequencies. Single-layer solenoids, and condensers, should show sensibly the same values at radio as at audio frequencies. The double-layer solenoid measured in Exp. 10 will probably act like a condenser at radio frequencies. Even if it is found to be an inductance, its value will have no relation to the value measured at 1000 cycles. And if it is measured at several radio frequencies it will probably act like a coil at some frequencies and a condenser at others.

In Exp. 18, an oscillation transformer is to be used in the operation of a simple spark transmitter. Find the transformer you are to use and measure the self-induction of its coils, for several positions of the clips, and measure the mutual induction of the two complete coils for several separations, from the closest coupling to the weakest available. These values of L and M will permit the construction of a coupling curve which will be necessary in carrying out Exp. 18.

EXPERIMENT 17

Object.—To measure the natural frequency of an antenna using buzzer excitation and an aperiodic listening circuit. To measure the capacity and self-induction of an antenna. To find the effect of the number of wires on the antenna capacity. To find the effect of a series inductance, a series condenser, and a condenser shunting a loading coil in the base of the antenna.

Analysis.—The actual antenna used either at a transmitting station or receiving station is not generally a very simple electrical conductor; in this first study of antenna theory we shall consider the antenna as a simple vertical wire having the same inductance and capacity per centimeter throughout its entire length, because it is only in an antenna of this kind that simple conclusions can easily be reached.

The vertical wire is to be regarded as one plate of a condenser, and the earth as the other. When a source of high-frequency a.c. power is connected, one side to the antenna and one to the earth, alternating currents flow into the antenna to charge it alternatively positive and negative. Of course the current at the extreme end (top) of the antenna must evidently be zero, and it follows therefore that the current is different (at a given instant) at different parts of the antenna. If the frequency is low this current change is nearly linear, as indicated in Fig. 86; and the voltage of the antenna, with the earth regarded as zero potential, is uniform as indicated by the voltage curve of Fig. 86. If we use the expression $I = 2\pi fCE$ and solve for C, the value obtained will be the capacity of the antenna when all parts of the antenna are charged to the same voltage. The current drawn from the alternator is of course a leading one. Now as the frequency is increased it will be found that at a certain frequency (called the natural frequency of the antenna) the current taken from the alternator is in phase with the impressed voltage; the power factor of the antenna is unity. An investigation of the voltage and current distribution for this condition shows them both to have sinusoidal form, as in Fig. 87. Each curve is one-quarter of a sine wave, their space distribution being 90° apart.

The velocity of travel of an electrical disturbance along a wire like this antenna is practically 3×10^8 meters per second, the velocity of light. The frequency impressed upon the antenna of Fig. 87 sets up one-quarter of a wave length, so we can say that the antenna is radiating for this condition a wave four times as long as the antenna is high. Thus if the antenna of Fig. 87 is 50 meters high, the wave length radiated is 200 meters and the frequency of the current must be $(3 \times 10^8)/200 = 1.5 \times 10^6$ cycles per second.

Now any circuit which we know has capacity, and yet shows unity power factor must have an inductance, and must have as much magnetic energy stored as $LI^2/2$, as it has electrical energy stored as $CE^2/2$. The inductance of the antenna is of course due to the magnetic field which the antenna current sets up in the surrounding region; as there is no current at the top of the antenna then there can be no magnetic field there, and also it follows from Fig. 87 that the magnetic field must

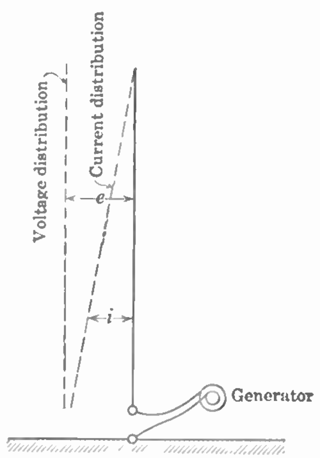


FIG. 86.

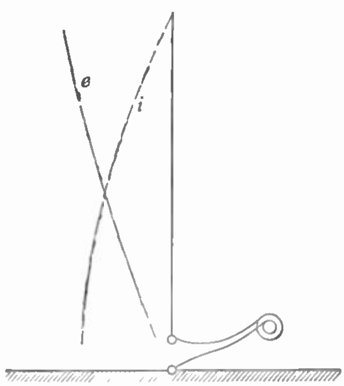


FIG. 87.

be a maximum near the base of the antenna, decreasing sinusoidally toward the top.

The inductance of the antenna is calculated by measuring the resonance frequency (¼ wave length oscillation) for the condition of Fig. 87; with the capacity obtained from the condition shown in Fig. 86, and the formula for resonance, $f = \frac{1}{2\pi\sqrt{LC}}$, L is then calculated. Evidently this is a fictitious value of L, because the capacity of the antenna for the voltage distribution of Fig. 87 is considerably different from the capacity with the voltage distribution of Fig. 86.

In spite of this discrepancy in reasoning, however, the values of capacity and inductance obtained by the method outlined above are known as the antenna capacity and inductance, respectively. As generally used, the voltage distribution of the antenna is more like that of Fig. 86 than that of Fig. 87, so that the capacity value is reasonably correct; the inductance of the antenna is generally small compared to the inductance added in series with the antenna, so that even if there is a considerable error in the assumed value of antenna inductance, but little error is caused in the final result.

In the case of a simple inverted L antenna (Fig. 88), the natural wave length is four times the total length from earth to extreme end. If the top part of the antenna consists of several wires connected together in parallel, the natural wave length may be as much as six times this length. In the case of a T antenna the natural wave length is from four to six times the distance from ground up the down lead and out the antenna to one end of the top part.

FIG. 88.

The capacities of antennas used today, for transmitting stations, range from 0.0005 to 0.005 microfarad; the capacity of the small antenna used for a receiving station may be only 0.0002 microfarad. The inductance of the small receiving antenna is a few microhenrys, say from 10 to 30, depending upon the length.

An antenna intercepts the broadcast wave, and a voltage is set up in the antenna equal to the " microvolts per meter " strength of the wave multiplied by the *effective* height of the antenna in meters. (The effective height is generally about 0.8 the actual height.) But little current will flow in the antenna circuit, however, unless the antenna is tuned to the frequency of the wave itself. Suppose an antenna 10 meters high having $C = 0.00015\mu f$ and $L = 10\mu h$ and R (effective) of 5 ohms, is excited by a " one millivolt per meter " wave from a 1000 kilocycle transmitter. How much current will flow? $X_c = 1062$, $X_L = 62$, $X = 1000$, $Z \cong 1000$, so $I = E/1000 = 0.001 \times 10/1000 = 0.00001$ ampere.

If, however, we put an inductance in series with the base of the antenna such that its reactance combined with the inductive reactance of the antenna just neutralizes the capacity reactance of the antenna, the coil having 5 ohms resistance, then the impedance of the antenna circuit will be merely its resistance, or 10 ohms. The current then will be 0.001 ampere, or 100 times as great as for the untuned antenna.

The tuning of an antenna therefore is a very important condition to be fulfilled in using it for picking up, or radiating, energy, and this constitutes the main object of this experiment.

The top part of an antenna, especially that of a transmitting station, is generally made up of several wires all connected in parallel. Suppose five wires are used, will the capacity be five times as much as the capacity of a single-wire antenna? Sometimes two or more down leads are used in parallel; is the inductance of such an antenna the same as if a single down lead were used? The answering of these two questions forms another part of this test.

To make the proper measurements for anwering the two above questions, it is convenient to have a flat-top antenna with, say, five wires. Three wires (the two outside ones and center one) are connected together and to one down lead; the other two are connected together and to another down lead about 2 feet from the first, and parallel to it. The inductance and capacity of each of the three antennas (two-wire, three-wire, and five-wire) are then measured, and suitable comparisons will give at least qualitative answers to the questions asked.

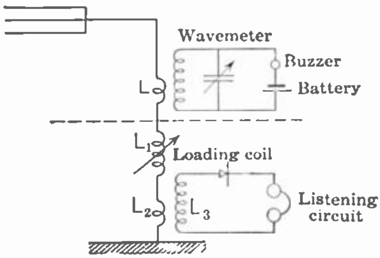


FIG. 89.

The antenna is to be excited by a wave meter, buzzer excited. An aperiodic listening circuit is loosely coupled to the antenna, and by its indications resonance between wave meter and antenna can be obtained. The arrangement is as shown in Fig. 89. A coil of a very few turns L, is coupled to the wave meter, which is generating high-frequency current by the action of its buzzer. An adjustable inductance L_1, of range up to about $1000\mu h$, is in series with the down lead and also a coil of a few turns L_2. This coil is coupled to the listening circuit consisting of coil L_3 in series with a detector and phones. The dashed line across the center of the figure is to indicate that there must be no appreciable coupling between the wave meter and listening circuit; whatever energy gets into the listening circuit from the wave meter must come through the antenna.

The coil L_3 should be variable to some extent; the secondary coil of the old-fashioned "loose coupler" is satisfactory. The coils L and L_2 should be of as small inductance as can be used. When they are too small the listener has difficulty in ascertaining the resonant frequency of the antenna; when they are too large, they upset to some extent the calculation of the antenna constants, because their inductance is unknown, and neglected in the following procedure.

The experiment is carried out with the least trouble if the first measurement made is to find the resonant frequency of the antenna with L_1 set at its largest value. For properly carrying out this test, on the average small antenna, a calibrated variometer with a maximum inductance of 1000 to 1500 microhenrys is suitable for L_1. By neglecting the inductance of the antenna itself, as well as that of L and L_2, the resonant frequency of the antenna circuit is approximately calculated from the known value of L_1 and a reasonably assumed value of antenna capacity. (This will vary from $100\mu\mu f$ for a small receiving antenna to perhaps $600\mu\mu f$ for one having several wires in parallel in its elevated portion.) This rough approximation enables the experimenter to set the wave meter generator at approximately the correct frequency for resonance, without wasting time.

In this test, coils L and L_2 may be as large as perhaps five turns, of say 4 inches diameter, without appreciably affecting the accuracy of the result. It is convenient to make these coils by winding them loosely of flexible wire; the number of turns can then be easily altered. Of course, if the L_2–L_3 combination is a loose coupler, L_2 will be equipped with a switch by which the desired number of turns can readily be selected.

With tight coupling between the wave meter coil and L, and tight coupling between L_2 and L_3, the wave meter setting is changed until the signal is loudly heard in the phones. Because of the tight coupling used, it is likely that the setting for a loud signal in the phones is not well marked. By decreasing the coupling, both between wave meter and L, and between L_2 and L_3, the setting of the wave meter, for resonance, becomes more definite, and of course the signal generally becomes weaker.

With a little practice the student should be able to set the wave meter condenser on the resonant point within one division or better.

The listening circuit is supposed to be aperiodic, i.e., non-resonant, but it sometimes happens (especially if L_3 is a coil of many turns) that the *natural period of the coil* L_3 is the reason for the resonance showing. This is especially true when the value of L_1 being used is small; it is not likely to occur with large values of L_1. However, to test for this possibility, the number of turns in L_3 is changed somewhat and the wave meter is again adjusted for resonance. If this setting of the wave meter is the same as the previous one (with original value of L_3) the setting is actually the resonant frequency of the antenna. If, however, the setting of the wave meter is changed (for resonance) with the new value of L_3, then the frequency indicated by the wave meter is the resonant frequency of coil L_3 and not that of the antenna.

Having measured the wave length of the antenna with L_1 in series, the capacity of the antenna can be calculated from the relation

$\lambda = 1884\sqrt{L_1 C_{\text{ant}}}$. This formula neglects the inductance of L and L_2 as well as the inductance of the antenna. Provided that L and L_2 are kept as small as feasible and L_1 is of the order of 1000–1500 microhenrys, but little error occurs in this determination of the antenna capacity. If desired a small correction can be made to this value of capacity later in the experiment.

Now the value of L_1 is decreased in about five steps to its minimum value, reading the resonant wave length of the antenna circuit for each setting. Then another variometer is substituted for L_1, this new one having a maximum inductance about equal to the minimum inductance of former one. Get the resonant frequency of the antenna for about five values of this new inductance, the values being spaced approximately equally between its maximum and minimum inductances.

As the values of this added inductance are continually diminished, the error caused by the neglect of the inductances L and L_2 continually increases. Hence as the test progresses, the number of turns in L and L_2 should be reduced as much as feasible, keeping only enough turns to give a well-defined resonance setting of the wave meter.

Get a reading with the inductance L_1 taken out entirely, of course connecting together the two wires of the antenna circuit where the coil L_1 is taken away. This reading gives the *natural frequency* of the antenna, except for the slight error caused by L and L_2. It will be noticed that the resonance setting is not as well defined as was the case when large values of L_1 were used. As the amount of loading inductance, L_1, is diminished, the decrement of the antenna circuit increases, thus making the resonance broader (not so sharp).

From this natural frequency of the antenna and the previously calculated antenna capacity, the value of the *antenna inductance*, L_{ant}, can be calculated. If it is found that this is appreciable compared to the largest value of L_1 then this can be added to the value of L_1 and the antenna capacity recalculated from the formula

$$\lambda = 1884\sqrt{(L_1 + L_{\text{ant}})C_{\text{ant}}} \qquad . \quad . \quad . \quad . \quad . \quad (40)$$

With this new value of C_{ant} we can calculate a somewhat more accurate value of L_{ant}. The inductance of the antenna (including that of L and L_2) can also be obtained from the results of the experiment just carried out, by the help of a curve as shown in Fig. 90. By plotting the relation between the values of L_1 and (wave length)^2 a straight line curve is obtained which cuts the X axis with a negative intercept. This negative intercept, to the same scale as used for L_1, gives the antenna inductance.

With the maximum value of L_1 added, measure the resonant frequency for the two other antennas. If the five-wire antenna was used in the above test, then measure the resonant frequency for the two- and the three-wire antennas. Calculate the antenna capacity for each case.

With no added inductance, measure the natural frequency for these two other antennas. Calculate the antenna inductance for these two other antennas.

It should be found that the capacity goes up with the number of wires, but by no means is it proportional to the number of wires. Assuming two down leads, some distance apart, it should be found that the inductance of the three-wire antenna is slightly more than that of the two-wire antenna and the inductance of the five-wire antenna (the two down leads in parallel) is considerably less than that of the three-wire, and possibly less than that of the two-wire antenna.

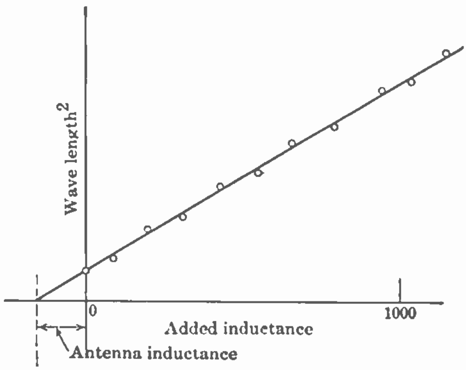


FIG. 90.

In place of L_1 (Fig. 89), insert a calibrated condenser of maximum capacity several times as much as the antenna capacity. Measure the resonant frequency with this condenser set at its maximum value; this frequency is somewhat greater than the natural frequency. It is also more difficult to obtain because, as the resonant frequency of the antenna circuit increases, its resistance, and decrement, both increase; hence the amount of power supplied to the antenna by the wave meter generator diminishes, and the resonance becomes less definite.

Try to get values of resonant frequency for several decreasing values of the added series condenser; it will be found that as the value of this capacity approaches zero, the resonant frequency does not approach infinity, as might be supposed, but approaches a frequency just twice the natural frequency.

In case too much difficulty is experienced in getting settings to show this effect of the series condenser, add a coil of a few hundred microhenrys in series with the antenna (as well as the series condenser) and leave this coil fixed as the series condenser is varied. A curve somewhat similar to the desired curve will now be obtained, showing that as the capacity of the series condenser is diminished, the resonant frequency continually increases. All frequencies will, however, be lower than those which would have been obtained if no added inductance had been used in the antenna circuit.

Put the variable inductance L_1 (formerly used) in series with the antenna and set at its maximum value. Connect the variable condenser in shunt with the coil, and find how the resonant frequency of the antenna circuit varies as this condenser is varied from zero to its maximum value. Get about six points.

Plot your results in suitable curves, using for the abscissae the factor which was varied, inductance or capacity, as the case may be.

EXPERIMENT 18

Object.—To study the operation of a small spark transmitting set. Tuning closed and antenna circuits. Effect of coupling on frequency of radiated power. Conditions for maximum radiation at a specified frequency. Effect of antenna resistance on proper coupling. Operation of a quenched spark.

Analysis.—In radio communication as carried on today, spark telegraphy plays a minor role; a study of the operation of a spark set is, however, worth while because of the theory involved, which theory is applicable to all shock-excited oscillations.

In Exp. 16 we utilized the high-frequency oscillations which are set up when the energy which is stored up in the magnetic field (as $LI^2/2$) is emptied into a condenser. In the present experiment we study the effect of a condenser, in which electric energy is stored, being connected to a coil. As before, the stored energy sets up oscillatory currents. The oscillatory power so generated has had a very wide application in radio, practically all transmitting sets having operated in this fashion up to a few years ago. Even now, in the merchant marine, there are thousands of radio transmitters which obtain the high-frequency power, required for radiation, by the sudden discharge of a condenser through a suitable coil.

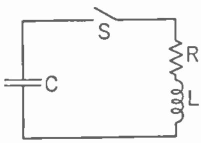


FIG. 91.

If a condenser C is charged to a voltage E, the stored energy is $CE^2/2$. Now if such a condenser is connected to a coil of inductance L and resistance R (as in Fig. 91), when the switch is closed oscillatory current will flow in the circuit, the current being given by the equation

$$i = E\sqrt{\frac{C}{L}}\epsilon^{-\frac{Rt}{2L}} \sin 2\pi ft \quad . \; . \; . \; . \; . \; . \; . \quad (41)$$

in which $f = \dfrac{1}{2\pi\sqrt{LC}}$. If we write $\delta = \dfrac{R}{2fL}$ and let n represent the number of cycles of current that have occurred since the switch was closed, the equation becomes

$$i = E\sqrt{\frac{C}{L}}\epsilon^{-n\delta} \sin 2\pi ft \quad . \; . \; . \; . \; . \; . \; . \; . \; . \quad (42)$$

The term $\epsilon^{-n\delta}$ continually decreases the value of the right-hand term; the function, $\sin 2\pi ft$, is zero when the switch is closed ($t = 0$) and rises to its maximum value (unity) one-quarter of a cycle later. The maximum current which occurs in the circuit is theoretically (no damping) given by $E\sqrt{\frac{C}{L}}$; we call it I_0. The actual maximum current is $I_0\epsilon^{-\delta/4}$, which in the average radio transmitter circuit is not greatly different from I_0. In a spark transmitter a spark gap takes the place of the switch of Fig. 91; the arrangement is shown in Fig. 92. The power of an alternator is stepped up in voltage to about 15,000 volts, and the voltage is impressed on condenser, C; the coil L of a few microhenrys serves as a connecting wire while the condenser is being charged.

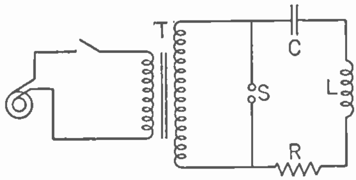


FIG. 92.

When the voltage reaches a certain value, this depending upon the length of the gap, the spark gap breaks down and oscillatory currents flow in the C–L–R–S circuit of Fig. 92. During this time the gap is conductive and practically short-circuits the secondary of transformer T. Evidently this transformer must be one with high magnetic leakage, or the alternator would feed so much energy into the gap as to burn up its electrodes. Generally T is either an open magnetic circuit transformer (lots of leakage) or high reactance is introduced between the alternator and spark gap by other means.

The frequency of oscillations in the C–L–R–S circuit is determined by $f = \frac{1}{2\pi\sqrt{LC}}$ and the rate of decay by the factor $\epsilon^{-n\delta}$. Now in this circuit, as necessarily constructed in a powerful transmitter, this decay term is large; the decrement may be several tenths.

This is caused by the small value of inductance permissible in the circuit. For a given power rating of transmitter a given amount of energy must be stored in the condenser before each spark. If there are N sparks per second, if the condenser will stand a voltage E safely, and has a capacity C, the energy transformed into high frequency oscillatory power, each second, is given by

$$\text{Power} = \frac{NCE^2}{2}. \quad \quad (43)$$

If there are 1000 sparks per second, and the condensers will stand safely 10,000 volts, we need a condenser of 0.02 microfarad per KW.

of high-frequency power. A 2-KW. set would then require a condenser of 0.04 microfarad.

If the frequency to be generated is 500,000 cycles (standard for the merchant marine) the amount of inductance required is only 2.5 microhenrys. Now it is impossible to build such a circuit with a low decrement; the inductance is too small.

The energy in this circuit is then used to set up oscillations in another circuit in which a low capacity and high indutance are used; here the decrement may be made reasonably low. This second circuit is the antenna circuit of the ordinary ships set, as in Fig. 93. To get the energy from the local closed circuit S–C–L–R into the antenna this must be tuned to the same frequency as the closed circuit; that is, $C_1L_1 = CL$. Now C_1 is only about 0.001, so that L_1 will be twenty times as large as L and the decrement may be kept below 0.2, the maximum permitted by law.

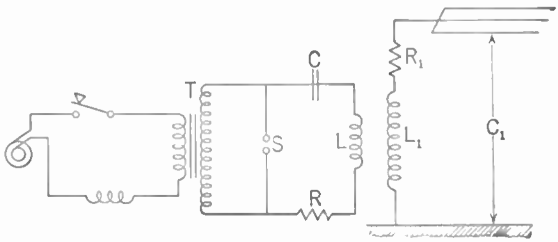


FIG. 93.

As long as S stays closed there are two tuned coupled circuits acting here, so that the theory analyzed in Exp. 9 must apply. That experiment showed that there were two resonant frequencies for such a combination, and neither of these two frequencies was the one for which both circuits are tuned. The high-frequency oscillatory energy will divide nearly equally between these two resonant frequencies so there will be, in the arrangement of Fig. 93, two high-frequency currents in each of the coupled circuits. One frequency will be higher than the frequency for which each circuit is tuned, and the other will be lower. As shown in Exp. 9, the separation of these two frequencies depends entirely upon the coefficient of magnetic coupling of the two circuits, the tighter the coupling the farther apart the two frequencies.

The two frequencies are (by eqs. 30 and 31)

$$f' = \frac{f}{\sqrt{1+k}} \quad \text{and} \quad f'' = \frac{f}{\sqrt{1-k}}$$

so that

$$k = \frac{f'' - f'}{f} \qquad (44)$$

An antenna which sends out two frequencies, neither of them being the frequency for which the circuit is tuned, is evidently of but little value. The phenomenon is, however, an interesting one and is to be investigated in the laboratory.

A hot wire meter in the antenna circuit will read the effective value of the series of damped sine waves, perhaps a thousand per second. Its reading will, of course, be but a small fraction of the maximum current I_0 in the circuit. In a typical case I_0 was 260 amperes but the reading of the hot wire ammeter in the circuit was only 15 amperes.

Now it will be found that the reading of the antenna ammeter varies with the amount of coupling between the two circuits; as the coupling is decreased from its tightest value the antenna ammeter decreases slowly at first and then quite rapidly. It might then be thought that tight coupling should be used between the antenna and closed circuit, because here the antenna ammeter gives its greatest reading. Such is far from the fact, as the test next to be described will show.

A wave meter, tuned to the frequency which the set is supposed to generate, is coupled loosely to the antenna circuit. The reading of its hot wire ammeter corresponds to the response which would be received by a distant receiver circuit tuned to the frequency which the transmitting antenna is supposed to send out.

Now this curious result will be obtained. With tightest coupling the wave meter reads practically zero; as the coupling is decreased the antenna ammeter gradually decreases its reading, and the wave meter ammeter reading increases! This effect will continue, antenna ammeter decreasing, wave meter ammeter increasing until a critical coupling is reached, beyond which the wave meter will rapidly fall. This critical coupling, it will be noticed, is that value at which the antenna ammeter begins its rapid decrease, after having held a nearly constant value as the coupling decreased; the effect is shown in Fig. 94. As the wave meter reading corresponds to the signal which would be received by a distant listening station, it is evident that this critical coupling is the proper one for the transmitting station to use. Experiment will show that the critical coupling depends upon the resistance of the antenna circuit, the critical coupling increasing as the antenna resistance increases.

Now it is advantageous, in many ways, to use a low-resistance antenna. The lower the decrement of the antenna circuit the sharper is the tuning of its signal in the distant receiving set, and hence the signal is freer from interference in the listener's circuit; also it causes less interference to those listening for other signals on neighboring frequencies.

But we have seen that a low-resistance antenna requires low coupling at the transmitting circuit, and low coupling here means that the high-frequency energy passes slowly from the closed circuit to the antenna circuit. But this closed circuit has a very high decrement, as explained before, and so the circuit rapidly uses up any energy oscillating there. Hence weak coupling at the transmitter really means that most of the high-frequency energy obtained from the condenser discharge is used up as heat in the closed circuit and so is of no use for communication purposes.

The radio engineer therefore has the task of using a low-resistance antenna circuit, a tight coupling between closed and antenna circuits, and yet have the antenna receive a maximum energy at the desired frequency. This requirement led to the development of a special type of spark gap, called the quenched gap.

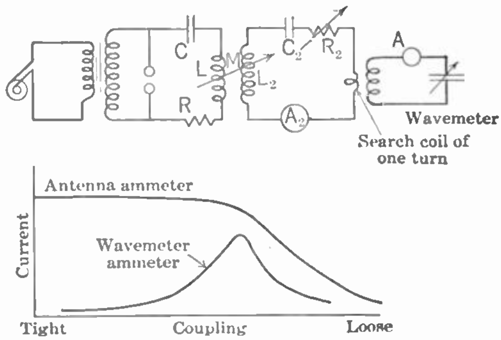


Fig. 94.

It will be noticed that the spark gap is in series with the closed circuit; if the gap is open no current can flow and the circuit does not exist. Then there is only one oscillating circuit (the antenna circuit) and this will oscillate freely at its own natural frequency.

The oscillatory energy starts in the closed circuit. It is originally all stored in the condenser C. Just as soon as oscillations start, the energy begins to flow across into the antenna circuit and after a few cycles (equal to $\frac{1}{2}k$) the energy has all been transferred to the antenna. Now if a set is equipped with a quenched gap the " closed " circuit will open at this instant, and the energy is left to oscillate in the antenna at its own frequency and decrement.

A quenched gap in good condition will restore its insulation, that is, change from a conducting gap to a non-conducting gap, in about 0.000001 second. Such a gap is made up of a series (generally ten or more) of short gaps between heavy copper discs. Each gap is about 0.008 inch long and the faces of the copper discs, from which the sparks jump, are made of silver, or else pure copper in the cheaper ones. Each gap is closed around its periphery by an insulating gasket, so that the gap is really an air-tight chamber. As the oxygen rapidly disappears, the spark normally takes place in nitrogen. This non-oxydizing gas permits the sparking surfaces to stay bright and clean, in spite of the millions of sparks that jump across the gap.

A quenched gap is easily spoiled. If the gaskets leak air the spark surfaces soon become covered with oxide and the gap will not quench. If the gap overheats it is also spoiled, and this overheating occurs if the coupling between the closed circuit and antenna is either too tight or too loose. With the normal gap a coupling between 20 and 25 per cent is necessary to retain its efficient operation.

In Fig. 95 are shown the currents in closed and open circuits when using an ordinary open (non-quenching) gap and when using a quenched gap functioning normally. With a quenched gap, much more power, at the desired frequency, can be obtained than with an open gap, assuming the same power supplied to the condenser of the closed circuit for both cases.

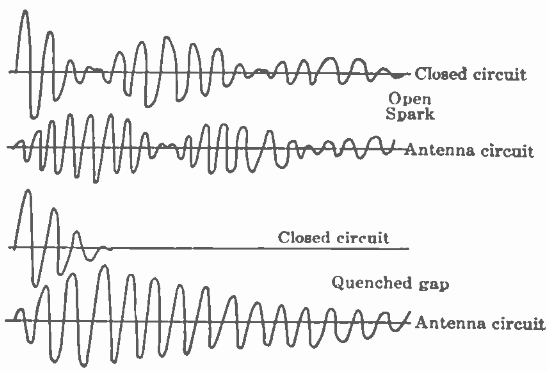


Fig. 95.

Using an open gap adjust the closed circuit of a spark transmitter to oscillate at any convenient frequency, while the spark gap is inspected and adjusted if necessary. The gap should be clean and adjusted for such a separation that the voltage required to break down the gap will not over-stress the condenser in the circuit. Have an ammeter in the oscillatory circuit to indicate the current therein; it will be noticed that unless the spark is white and snappy there is practically no high-frequency power generated.

A bluish, roaring spark (really an arc) shows that the gap never really opens—the power supply sends sufficient power to it so that it maintains its conductive condition. This being the case, the secondary winding of the transformer is continually short-circuited, so that there is never available a high voltage for properly charging the condenser. And, of course, if the condenser never gets charged to the high voltage (several thousand volts) there is no high-frequency power generated and so the ammeter in the high-frequency ammeter does not read.

To get a snappy spark, it may be necessary to get more reactance between the power supply and the spark; a choke coil may possibly have to be used in series with the primary of the transformer. Having the spark operating properly, find the relation between oscillatory current and amount of capacity, with fixed spark-gap length, and the relation between oscillatory current and gap length, for a fixed amount of capacity.

With proper amount of capacity and spark-gap length, vary the inductance one turn at a time and read the wave length for each adjustment.

One clip of the oscillatory circuit should be left connected to the outer end of the spiral, and the other clip moved in toward the center of the spiral. The outer turns of the spirals permit tighter coupling than the inside turns, and later in the test tight coupling is called for; if inside turns of the spiral were being used it could not be obtained.

Of course in using the wave meter in this test the hot wire ammeter, rather than crystal and phone, is used to detect resonance. Couple the wave meter to the coil of the circuit as loosely as possible so that a very small deflection is obtained on the ammeter of the wave meter. A small deflection shows the wave length just as well as a full scale reading, and with loose coupling the wave meter ammeter is less likely to be burned out by an accidental increase in coupling. It will be noticed in this test that a very slight change in angular position of the wave meter coil may result in large changes in its ammeter reading.

Set the clip on the inductance at the right place to get the wave length specified by the instructor, using outer turns of spiral. With the secondary coil of the oscillation transformer connected to a suitable condenser, ammeter, a small search coil and a variable non-inductive resistance in series, tune this circuit to the frequency of the primary. This is done by using loose coupling, small added resistance, and varying the inductance of the secondary (using outer turns of the spiral) until the ammeter indicates resonance.

The small search coil is for the purpose of coupling the wave meter to the secondary circuit in such a way that as the secondary coil is shifted (to change transformer coupling) the wave meter coupling is not changed. The search coil is conveniently made of one or two turns of flexible wire, the turns about 3 inches in diameter. This coil is in series with the secondary circuit; it is placed on the table and held in place by a clamp of some kind. The wave meter is placed so that it is coupled to the secondary circuit only through this search coil.

At this stage of the test it is to be noticed that the tuning of the circuit can be seriously affected by moving any of the flexible wires used in making up the circuits. There should be as few of these as possible, and they should be so arranged that they are disturbed as little as possible when carrying out the tests called for below.

With no resistance added in the secondary circuit and tightest coupling, spark operating steadily, enough so that the ammeter in the secondary circuit shows a reasonably steady reading, wave meter fixed at a coupling which gives reasonable readings on its ammeter take a set of readings between wave meter setting, and reading of its ammeter. This shows the "energy distribution" curve; it shows how the high-frequency energy is distributed among the various frequencies in the region of the tuned frequency. (Of course all the energy should preferably be concentrated in the frequency for which the circuit is tuned.) Repeat this for two other values of transformer coupling, medium and loose, keeping all other conditions the same.

With medium coupling take a series of readings between secondary current and secondary added resistance. (The rheostat should have about 50 ohms resistance, maximum.) Take care that the variation in secondary resistance does not mistune the secondary; if it does, retune by moving the clip on the secondary of the oscillation transformer. During the test have the wave meter set at the frequency for which the circuits have been adjusted. For about six values of added resistance read "antenna" current and ammeter of wave meter. (The secondary circuit in this test corresponds to the antenna of an actual spark transmitter.)

With no added resistance and very weak coupling get a "resonance curve" of the secondary current. Read the wave meter ammeter for various settings, keeping the coupling (wave meter to search coil) fixed. The shape of this resonance curve enables us to calculate the sum of the decrements of the wave meter and the secondary circuit. Knowing the wave meter decrement (from instructor) the circuit decrement is obtained as the difference between total decrement, from curve, and the wave meter decrement. If the decrement of the secondary circuit is known, its resistance can be calculated from resonant frequency and inductance. Thus the total secondary circuit resistance is now known.

With no added resistance, wave meter coupling to search coil constant, wave meter set for resonant frequency, take a series of readings between coupling, closed circuit current, secondary circuit current, and wave meter reading. Take readings for about eight different values of coupling.

Repeat the above run with about 10 ohms added to the secondary, and also with about 30 ohms added.

The wave meter coupling should be left the same for all three runs if possible, so that its readings give comparative "radiation" for the several conditions. (Of course there is practically no radiation from the laboratory circuit.)

Replace the open gap by a quenched gap, of about the same breakdown strength. With about 25 per cent coupling and about 5 ohms added resistance get an energy distribution curve of the secondary current. (Have the wave meter to search coil coupling the same as it was for the previous tests.) Read also the secondary current. Try the same test with tighter coupling and with looser coupling, say 40 per cent and 10 per cent. As the gap may heat up and spoil under these conditions get readings quickly and take off power after readings are obtained.

EXPERIMENT 19

Object.—Study of electron emission from various kinds of cathodes. Effect of space charge and anode voltage on the thermionic current.

Analysis.—We regard a conductor as being made up of an aggregate of atoms and electrons, both of which are in a state of haphazard motion, except at absolute zero temperature. Because of their relatively small mass, and the equipartition of energy between atoms and electrons, the velocity of the electrons is very high, even at room temperature.

The atoms of a metal tend to separate from each other as the temperature is raised, or we may say the metal tends to evaporate. Now it seems likely that if, when the atoms acquire a high velocity, they are able to break through the surface tension of the metal, the electrons can do the same thing, hence we get the idea of electrons evaporating. This evaporation of electrons will take place at lower temperature than that of the atoms of the metal itself because of the higher average velocity of the electrons.

Richardson predicted the evaporation of electrons would take place according to the equation

$$N = A\sqrt{T}\epsilon^{-\frac{a}{2T}}. \qquad (45)$$

N = number electrons evaporating per second per square centimeter of surface

T = absolute temperature, Centigrade
a = latent heat of evaporation
A = a constant

Later Dushman found that the evaporation took place nearly in accordance with the relation

$$N = AT^2\epsilon^{\phi_0 e/kT} \qquad (46)$$

in which A is a constant, ϕ_0 = Richardson's "work function," k is Boltzmann's constant, and e is the charge on an electron.

The work function ϕ_0 expresses the energy it takes to abstract unit electric charge from the metal; that is, it is a measure of the effort of the metal to keep its electrons within the boundary of its surfaces. This has been measured for some metals, and is found to be 4.53 volts for tungsten, 4.40 for tantalum, 4.31 for molybdenum, 2.94 for thorium, and 2.24 for calcium.

To be useful as a source of evaporated electrons it is evident that a metal must neither melt nor evaporate at the temperature required for electron evaporation. This consideration limits the electron-emitting metals to a comparatively small list—tungsten, thorium and some refractory oxides, notably barium and strontium.

Until very precise methods of experimentation had been developed it was a question (even as late as 1910) whether electrons actually did evaporate from metals at all. It is now known that almost imperceptible amounts of certain gases, in contact with the heated metal from which the electrons are expected, will so contaminate the surface that electrons cannot break through. The effect is much the same as occurs when water is covered with a very thin layer of oil; water cannot evaporate through such a layer, no matter how thin it is. Traces of water vapor in contact with hot platinum stop electron evaporation almost completely; Langmuir found that the surface of tungsten could be completely spoiled, as far as electron evaporation was concerned, if a pressure of only 0.000001 mm. of oxygen was in contact with the hot tungsten.

Some idea of the velocity required before an electron can break through the surface of a metal can easily be obtained. The energy abstracted from the electron as it breaks through the "surface tension" of the metal is equal to the electric charge in the electron multiplied by the surface work, measured in volts. An electron, inside the metal, must approach the surface with such velocity that its kinetic energy is at least equal to that demanded by the surface work. So we put

$$Ve = \tfrac{1}{2}mv^2 \quad \text{. . . . .} \quad (47)$$

where V = surface work
e = charge on electron
m = electron's mass
v = required velocity.

V, e, and m being known, v can be calculated. For tungsten this velocity proves to be somewhat greater than 10^8 cm. per sec., about 700 miles per second.

Any electron having this critical velocity would undoubtedly fall immediately back into the tungsten; only those electrons having greater velocity than this emerge from the tungsten with a residual velocity, which will carry them away from the surface of the tungsten and so make them available for producing current effects.

The filament of the ordinary incandescent lamp is of tungsten; when carrying its rated current such a filament has a temperature of about 2400° K. (K stands for Kelvin; the scale is the absolute Centigrade scale.) The number of electrons breaking through the surface of such a filament can be calculated from eq. (46); of the number breaking through only 1 per cent have a residual velocity (after coming through the surface of the tungsten) of 2.5×10^7 cm. per sec., and only 1×10^{-6} per cent have a velocity as high as 4×10^7 cm. per sec. It will be recollected that they all must have a velocity (inside the metal) of at least 10^8 cm per sec., to break through; it is then evident that they lose nearly all their energy in breaking through the surface tension of the tungsten.

A little thought shows at once that as a source of evaporated electrons we should use that type of surface which emits electrons copiously at a low temperature. Whatever heat is supplied for heating the filament, from which the electrons are coming, is evidently wasted, in so far as evaporated electrons are concerned, so this heat should be kept as small as possible. In fact the usefulness of an electron-emitting surface is sensibly measured in terms of "amperes per watt." A filament which gives us an emission equivalent to 1 ampere of current, with a heat expenditure in the filament of 10 watts, is quite evidently superior to one which requires 100 watts for the same emission.

As pointed out before, the surface work of thorium is only 2.94 volts whereas that for tungsten is 4.53 volts. It therefore requires only about two-thirds as much power to get electrons out of thorium as out of tungsten, so we would naturally use a thorium filament in preference to one of tungsten. However, thorium evaporates very rapidly even at the low temperature required for electron emission, hence it cannot be used in this fashion. However, it is found that tungsten may be coated with an extremely thin layer of thorium (only one atom thick) and the electron emission is the same as it would be from a thorium filament, and furthermore this one atom deep layer of thorium does not evaporate until the temperature greatly exceeds that at which a pure thorium filament evaporates. The adhesion between the thorium atoms and the underlying tungsten atoms is greater than that between two layers of thorium atoms.

This so-called thoriated-tungsten is made very simply. Thoria (thorium oxide) is present in tungsten as ordinarily reduced from its ores. If $\frac{1}{2}$ per cent of thoria is present there is sufficient thorium to repeatedly give the tungsten a thorium coating, as this is spoiled for some reason or other.

The tungsten filament is first flashed at a dazzling temperature (2800° K) to clean its surface thoroughly. It is then held at a temperature of 2100° K for some hours; the thoria reduces to thorium and diffuses to the surface. This thorium-coated tungsten, at 2100° K, gives 100,000 times as much electron emission as pure tungsten.

This process of covering of the tungsten with its thorium layer is styled "activation" of the tungsten. When activated at 2000° K a certain filament gave a final emission of 0.001 ampere, and reached 90 per cent of this emission in three hours; held at 2050° K the final emission was 0.00074 ampere, and 90 per cent of this was reached in one hour; held at 2150° K the final emission was 0.00025 ampere and 90 per cent of this was reached in 30 minutes. It is then evident that the activation of thoriated tungsten is a rather critical process.

If now a properly activated filament is held at 2250° K it rapidly loses its thorium layer, and in a short time the emission drops to practically zero. Thus a thoriated tungsten filament must be operated very closely to its rated temperature, or it may spoil. Although spoiled, as far as electron emission is concerned, it lights up properly so that one has no idea it has been spoiled. Only an emission test shows this.

A thoriated filament may also be spoiled even when operating at normal temperature. If it is in a poorly evacuated bulb, and the gas becomes ionized when the filament is negative (with respect to the other electrode) the positive ions of the dissociated gas bombard the filament. A very short period of such bombardment completely spoils the thorium layer.

The reactivation process for a spoiled thoriated filament is given by the manufacturer as follows: A 3-volt filament should be operated at 10 volts for 30 seconds to rid the surface of contaminations and then at 4.5 volts for 10 minutes. During this treatment there must be no voltage on the grid or plate of this tube. Correspondingly a 6-volt filament should be operated at 15 volts for 1 minute and then at 7.5 volts for 10 minutes. This reactivation process can be carried out several times before the supply of thoria in the filament is used up.

The oxide-coated filament is generally made from a very thin ribbon of platinum nickel alloy. A mixture of barium and strontium oxides is baked onto the filament by any one of several processes, giving a layer of oxide of about 2 milligrams per sq. cm. of filament. This form of filament is more efficient than any other, in so far as emission is concerned. It will spoil if run at too high a temperature, because the oxide cracks away from the filament.

In Fig. 96 there are shown the emissivity characteristics of the three types of filaments discussed; logarithmic scales are used for emission and power consumption, and the coordinates are so drawn that the curves plotted between emission and filament power are straight lines. The

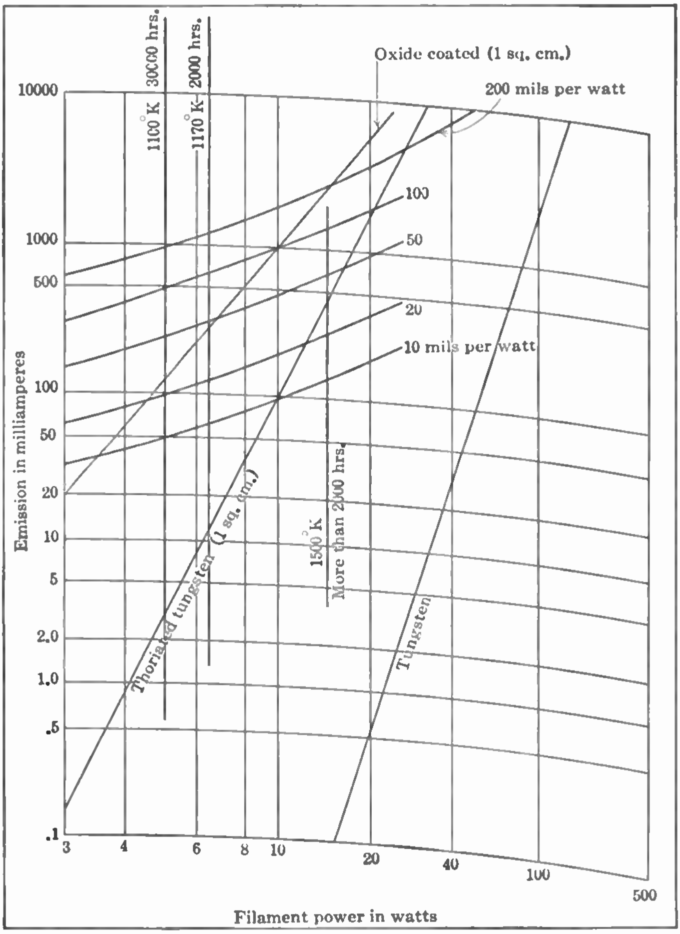


Fig. 96.

emission is given per square centimeter of filament, and the power is given in watts taken by this square centimeter of filament.

Suppose, e.g., that we have a tungsten filament of 0.3-mm. diameter and 10 cm. long; it will have about 1 sq. cm. of surface. If we supply 40 watts to this filament it will reach such a temperature that there will be 0.03 ampere of emission; if 60 watts are supplied to the filament it becomes much hotter (2370° K) and the emission increases to about 0.20 ampere. The filament will have about 700 hours' useful life at this temperature. If 100 watts are supplied to the filament it will rise in temperature to 2640° K and will last only 60 hours. However, its emission has increased to 2.0 amperes and the emission is 20 mils per watt.

Evidently then the emission per watt of filament power goes up rapidly; but the life of the filament is rapidly lowered. The proper operating temperature will therefore be fixed by the costs of filament power and cost of replacing the filament. The thoriated filament when run at such a temperature that the filament has 2000 hours of life (1500° K) gives an emission of 30 mils per watt, and the oxide-coated filament, run at 1170° K, has a life of 2000 hours and gives 50 mils of emission per watt of filament power.

To make measurements on the amount of electricity evaporated from filaments it is necessary to pull away from the filament all the electrons as soon as they evaporate, otherwise they fall back into the filament. Also it is necessary to have the electron-emitting surface in a highly evacuated bulb, to prevent the surrounding gas from contaminating the filament for one thing, and to permit the unimpeded motion of the electrons away from the filament. If there were appreciable gas surrounding the filament, the electrons, jumping through the filament surface as they evaporate, would rebound from the gas particles and re-enter the filament. Also some of the gas molecules would be split up (ionized), setting free electrons which would be confused with those evaporating from the filament.

Bulbs utilizing the phenomenon of electron evaporation are therefore highly evacuated, to about 10^{-6} mm. of mercury. Even at this high vacuum there are still about 10^{10} gas molecules per cu. cm., but this number is not generally harmful.

In Fig. 97 is a sketch of a filament, *F*, heated sufficiently by battery *A* to evaporate electrons *a*, *b*, *c*, etc. The plate *P*, reasonably close to the filament, is held at a positive potential with respect to the filament by the battery *B*. Any electrons flowing to the plate and then around the plate circuit through *B* and back to the filament constitute an electric current and will be registered by ammeter *C*.

The electrons flow in this circuit in the direction of the arrows, and at *O* join with those flowing from the *A* battery; past this point both plate and *A* battery current flow in the direction of the arrows to the filament. It follows that the current in the *N* end of the filament is greater than that at the *M* end by an amount equal to the current through ammeter *C*. This is, of course, as it should be; some of the electrons starting around the filament at *N*, escape (evaporate) and go over to the plate, so that there are fewer electrons passing point M in the filament than pass point N.

In the ordinary receiving tube the plate current is so small compared to the filament current that but little difference between the current at the two ends of the filament occurs; however, in a power tube, where the plate current may be 0.8 ampere and the filament current only 3.5 amperes, the current at one end of the filament may be 3.2 amperes and in the other 4.0 amperes. *This must be remembered.* It is brought out graphically in Fig. 98. The filament ammeter should always read the larger current; it should be in the negative leg of the filament with the plate return not connected between the ammeter and the filament, but on the other side of the ammeter.

If, with sufficient filament current to cause a perceptible electron evaporation, the voltage of the plate battery B (Fig. 97) is gradually increased from zero, the plate current will increase on a nearly parabolic curve for a time, then the increase becomes less rapid and finally the plate current becomes constant, independent of plate voltage. This is illustrated by curve Q in Fig. 99; with voltage from O to A this current increase is nearly parabolic; from A to B the slope decreases, and from B to C and beyond the current is constant.

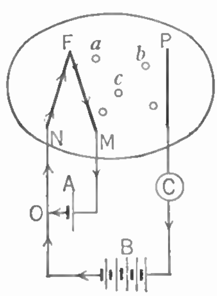


FIG. 97.

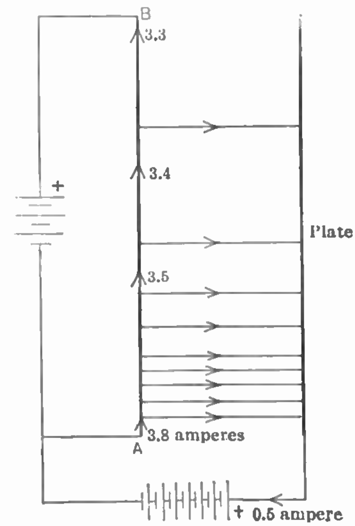


FIG. 98.

In discussions of this kind the term "plate voltage" always signifies the voltage of the plate *with respect to the negative end of the filament;* thus in Fig. 97 the "plate voltage" is the voltage to the negative end of the filament, that is in this case the voltage of the B battery. But if the B battery (negative terminal) had been connected to the positive terminal of the A battery, as it frequently is, then the plate voltage would be that of the A and B batteries combined.

Returning to the discussion of Fig. 99, we notice that the number of electrons *evaporating from the filament* must have been the same when getting all the points of curve Q, because the filament current, hence filament temperature, was held constant. Hence the electrons evaporating at this filament current are shown by the value OX, frequently called the saturation current, for this filament temperature.

Evidently for plate voltage less than OB, some of the evaporated electrons must have fallen back into the filament, and the reason is to be seen from Fig. 97. The electrons are distributed all through the space between the filament and plate, and these will tend to push back into the filament those electrons just emerging from the filament; these electrons which push back the others are called the "space charge" of the tube.

Until the plate reaches such a positive potential that its tractive force, on the electrons close to the filament, is great enough to overcome the repulsive effect of the space charge, some of the electrons will fall back into the filament. Under this condition the plate current is said to be *limited by the space charge.*

When the voltage exceeds the value OB (Fig. 99) and the plate current no longer increases as the plate voltage is increased, all the electrons which are evaporated come over to the plate—none fall back into the filament. The current is now *limited by emission.*

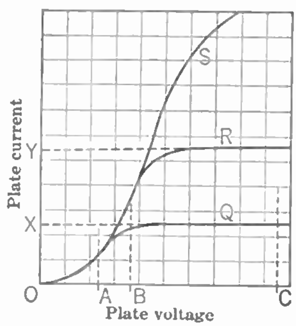


Fig. 99.

The plateau of plate current, shown flat in Fig. 99, is actually of this form in a highly evacuated tube with tungsten filament, or thoriated filament; with the oxide-coated filament saturation is never reached. The plate current continually rises with plate voltage, even though slowly. This effect probably has something to do with the rough surface of the filament, and possibly gas in the filament and tube plays a part. If now the filament current is increased (over the value it had in getting curve Q, Fig. 99, and the plate voltage again increased in steps, curve R is obtained. Again there is a parabolic portion and a plateau, but now the plateau begins at a higher plate voltage, and is higher. This means that the emission is greater; it is OY for this new value of filament current.

A further increase in filament current results in curve S; here the plateau is just beginning to appear, at the highest plate voltage used.

The emission, for this value of filament current, is greater than any value shown on this curve sheet.

In the curves of Fig. 99 the plate current is limited by space charge on the curved portions of the graphs, and by emission on the plateau. Suppose, now, that the plate voltage is held fixed, at some low value, and the filament current increased; the plate current follows the curve Q of Fig. 100. Up to a certain filament current no perceptible plate current flows; then it increases on a curve and finally becomes constant. For a given voltage the plate current does not increase with filament current after this reaches a critical value.

If the plate voltage is held at a higher value a similar curve is obtained, but the plateau is higher and only shows itself at a higher filament current. This is shown in the dotted curve of Fig. 100.

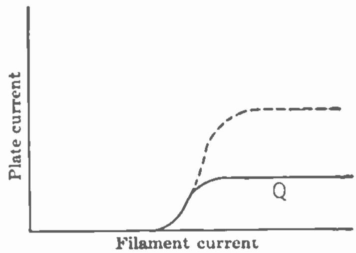


Fig. 100.

Now notice that on the curved parts of the graphs of Fig. 100 the current is limited by emission, and on the plateaus it is limited by space charge; these are exactly opposite to the conclusions reached when analyzing the similar-shaped curves of Fig. 99.

Using a small and a large tube of each type, get curves similar to those of Figs. 99 and 100 for oxide-coated, thoriated tungsten and pure tungsten tubes. In case two-electrode tubes of each kind are not available, three-electrode tubes may be used by connecting together the grid and plate, and treating the combination at the plate.

When carrying out the test use caution not to put on a higher voltage than the tube will stand. Of course, it appeals to any experimenter that the filament current must not exceed its rated value, but why must caution be exercised with respect to the plate voltage?

The power expended on the plate of the tube, which appears as heat, is equal to the product of the plate volts and plate current. If this exceeds a certain safe limit (of course different for different tubes) the plate becomes too hot and gas is emitted from the plate. This spoils the vacuum and the tube is ruined.

With the thoriated tungsten and oxide-coated filaments another trouble may occur owing to excessive bombardment of the filaments by the positive ions of the gas in the tube. Above a certain voltage this gas ionizes, and, even though there is not much gas present, the positive ions strike the filament with such momentum as to ruin it. In the thoriated filament all the thorium is knocked off, and in the oxide-coated one the bombardment is likely to concentrate on one spot, produce here an excessive temperature, and thus release enough gas from the filament to ruin the vacuum. It is well for the student to get from the instructor upper limits of filament current, plate current, and plate voltage for each of the tubes he tests.

Besides reading plate voltage and plate current in these tests read also filament power, so that the emission per watt of filament power can be computed.

Plot the curves of plate current *vs.* plate voltage on logarithmic cross-section paper, as well as upon ordinary cross-section paper. The exponent of the expression relating the two quantities in the expression

$$i = AE^x$$

may be measured from the logarithmic plot, as the slope of the line. If x is a constant in the above equation the plot will be a straight line; if x is different for different values of E (generally the case) the logarithmic plot will be curved. However, even if it is curved, the value of x for any value of E is equal to the slope of the curve at the specified value of E.

Note the comparative colors of the three types of filaments, when operating at rated filament power.

If time and apparatus permit, spoil a thoriated filament, by overheating, and then try to reactivate it according to the data given in the analysis above. This should be tried only with small amplifying or detector tubes; the process is not very successful with tubes above a few watts rating, because during the flashing part of the cycle, enough gas is generally driven off the filament to spoil the vacuum, so that when rated voltage is applied (after reactivation) the tube is likely to ionize, showing a light blue haze at first and then actually short-circuiting in the plate circuit if the plate voltage is maintained. This, of course, not only spoils the tube but also the ammeter in the plate circuit.

EXPERIMENT 20

Object.—Study of triodes, detectors, amplifiers, and power tubes. Calculation of amplification factor from curves and from tube dimensions.

Analysis.—It was explained in the previous experiment how the cloud of electrons between the filament and plate of a vacuum tube act as a hindrance to those near the filament; unless a comparatively high plate voltage is used the current to the plate is limited by this space charge.

As long as space charge limits the plate current, the plate current increases with the plate voltage according to the relation

$$i = kE^{3/2}/x^2. \qquad (48)$$

where E = voltage between cathode (filament) and plate
x = distance between cathode and plate

If x is in centimeters, E is in volts, cathode and plate are plane and parallel, this becomes

$$i = 2.33 \times 10^{-6}E^{3/2}/x^2 \text{ amp. per sq. cm. of plate} \qquad (49)$$

In case the plate is cylindrical in form and the cathode is a filament along its axis this becomes

$$i = 14.6 \times 10^{-6}E^{3/2}/r \text{ amp. per cm. length of plate} \qquad (50)$$

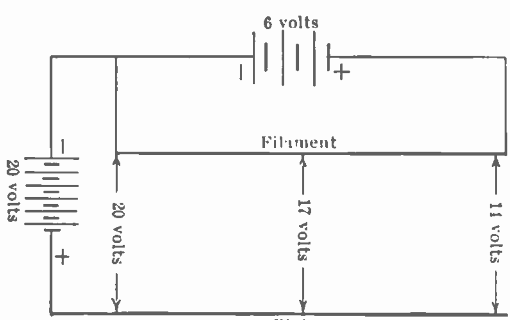


Fig. 101.

in which r = radius of cylinder in centimeters.

It is to be noticed that both of these expressions assume the same voltage between all parts of the cathode and anode (plate), but with the filamentary form of cathode this obviously is not true. In Fig. 101 this point is illustrated; a 6-volt battery is required to heat the filament and a plate battery of 20 volts is assumed. Evidently the potential gradient between anode and cathode varies greatly throughout the length of the filament. Under these conditions theory shows that the exponent of E in eqs. (49) and (50) should be much greater for low plate voltage than for high, and that the exponent is a variable, not constant.

If a unipotential cathode is employed the equations do give closely the form of plate current curve. Eq. (50), it should be pointed out, assumed the diameter of the cathode negligible compared to that of the anode.

Now if a mesh-like structure of conducting material is interposed between the cathode and anode, and its potential (with respect to the cathode) is varied, it will assist or oppose the space charge in controlling the plate current.

This extra electrode, called the grid, is often a ladder-like structure, of fine wires welded to rigid end wires; in other types of tube it is a helix of fine wires. The grid, as well as the plate and other metal parts inside the tube, is made of metal which will " out gas " readily, and which will hold its form when raised to a red heat. The alignment and spacing of cathode, grid, and anode must be held within very narrow limits if tubes are to have the same electric characteristics, plate current, grid control, etc.

A cross-section through two typical triodes (three-electrode tubes) is shown in Fig. 102. In a the grid helix is wound close, and the helix is long enough to reach past the ends of the filament, whereas in b the helix is much coarser and shorter.

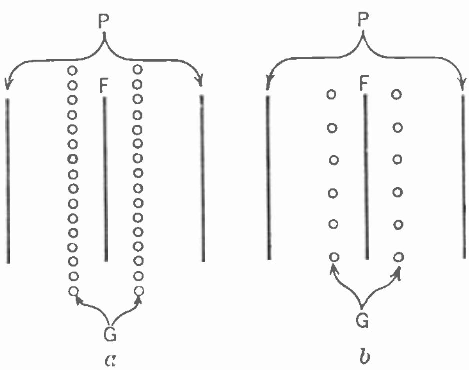


FIG. 102

The grid of a will control the plate current much more closely than will the grid of b.

In the triode the plate current will evidently depend (providing there is an emission from the cathode greater than is required for the plate current) upon both grid potential and plate potential. The equation for plate current is then

$$i = k(E_p + \mu E_g)^x \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (51)$$

It is necessary to put in the factor μ because it is not at once evident that 1 volt on the grid will have the same effect, on the plate current, as 1 volt on the plate; we may call μ the *relative effectiveness factor* of grid and plate potentials. The exponent is now written as x; it was stated previously that it was generally different for different voltages. Theoretically equal to $\frac{3}{2}$, the exponent is more nearly 2 for tubes as actually constructed.

The factor μ, for triodes as actually constructed, is greater than 1, in other words the grid voltage is more effective in controlling the plate current than is the plate voltage itself. Thus an increase in plate voltage (in a certain tube) of 1 volt might increase the plate current by 1 milliampere, whereas an increase of 1 volt in the grid potential (plate voltage remaining fixed) might increase the plate current by 10 milliamperes. The factor μ is therefore called the *amplification factor* of the tube. As it depends upon the geometry of the tube (size and spacing of grid wires, distance between grid and plate) it is generally called the geometrical amplification factor of the tube. It depends upon the various factors according to the equation

$$\mu = \frac{2\pi a n}{\log_e \frac{1}{2\pi r n}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (52)$$

in which a = distance between grid and plate
n = number of grid wires per cm
r = radius of grid wires

This formula is derived on the assumption that cathode, grid and plate are all infinite parallel planes; in the actual triodes a considerable departure from the value calculated from eq. (52) may be expected.

The potential of the anode with respect to the cathode is always positive. The cathode potential is always considered zero; in actual circuits the cathode is generally connected to earth.

The potential of the grid may be either positive or negative, but as used today the grid is practically always negative. If the grid is positive, some of the electrons on their way from the cathode to the anode will enter the grid and so constitute a *grid current*. However, if the grid is negative the electrons passing through it on their way to the plate are repelled by the grid wires, so that there is no grid current. Even when it is negative the grid still exerts its control over the plate current, so that the power output of the plate battery is controlled by the grid without this requiring any appreciable power. The triode is thus an amplifier, and this is the main service it performs today.

In Fig. 103 is shown a circuit for getting the characteristic curves of a triode. The battery E serves to heat the filament; the B battery serves to hold the plate at any desired positive potential;

the C battery, through the potentiometer M, serves to adjust the grid potential to any desired positive or negative value. The several ammeters A_f, A_b, A_c serve to read the currents of the various circuits. If another ammeter is inserted at A'_f, it will be found that this reading is less than that of A_f by an amount equal to the sum of A_b and A_c.

With a circuit arrangement as shown in Fig. 103 get curves of grid and plate currents for several types of commercial triodes. Get curves for normal plate voltage, 25 per cent over normal voltage and 25 per cent under normal voltage—or such other voltages as the instructor may specify. The grid potential variations should be sufficient, for each setting of plate voltage, to vary the plate current from somewhat greater than its rated current down to nearly zero. For many of the triodes used at present this means that no positive grid potential may be used.

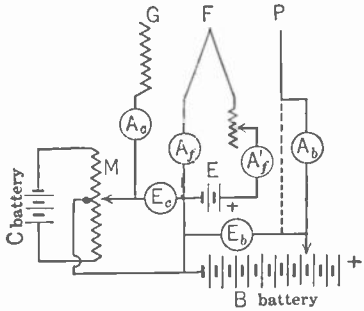


FIG. 103.

It will be noticed that the voltmeters are so connected that the respective ammeters do not read the current taken by the voltmeters. If the voltmeter E_b should have one of its terminals connected to the cathode and the other directly to the plate (as indicated by the dotted line), the ammeter A_b would read the voltmeter current plus the plate current; as the voltmeter current may be much greater than the plate current, this test, if so conducted, would give the characteristic of the voltmeter rather than that of the triode!

It is convenient to use for the meters A_c and V_c those with a center zero scale; it will then not be necessary to reverse meter connections during the test. Whereas there should, theoretically, be no grid current when the grid potential is negative, owing to imperfections in the tube some will generally flow—only a few microamperes, however. This current depends upon actual leakage between grid and cathode and upon the ionization of any gas which happens to be in the tube. When the grid is negative any positive gas ions on their way to the cathode may flow to the grid, thus giving grid current proportional to the amount of gas ionized. This form of grid current curve therefore is somewhat complex, depending upon the amount of gas ionized as well as the grid potential. The curve is used by triode manufacturers as a measure of the amount of gas left in the tube after the evacuation process has been completed. A typical curve is shown in Fig. 104.

For each triode tested there should be available a similar one with the glass bulb removed so that proper measurements can be made to compute the amplification factor from eq. (52).

From the plotted curves, of plate current *vs.* grid potential, the μ of the tube can be calculated by noticing how much the grid potential had to be varied to bring the plate current back to some specified value, after the plate voltage had been changed. Thus a plate current of 5 mils is obtained with $E_b = 150$, $E_c = -10$, and also with $E_b = 200$, $E_c = -20$, it is evident that a change of 50 volts in the plate potential is offset (in its effect on plate current) by a change of 10 volts in the grid potential. Thus 50 volts on the plate is "worth" only 10 volts on the grid so that the amplification factor is 5.

It will be found that the triodes available on the market have amplification factors between 2 and 50; the lower values are for the "output tubes" of radio sets and the higher values are for certain types of telephone repeaters.

If time permits get one or two plate current-grid voltage curves, for one of the tubes tested, with an added resistance in the plate circuit. The resistance should be about equal to the resistance of the plate circuit itself, that is, rated plate voltage (with zero grid voltage) divided by rated plate current. The plate circuit voltages used in these runs should be the same as those previously used so that a comparison of the curves may show directly the effect of external plate circuit resistance on the characteristic curves.

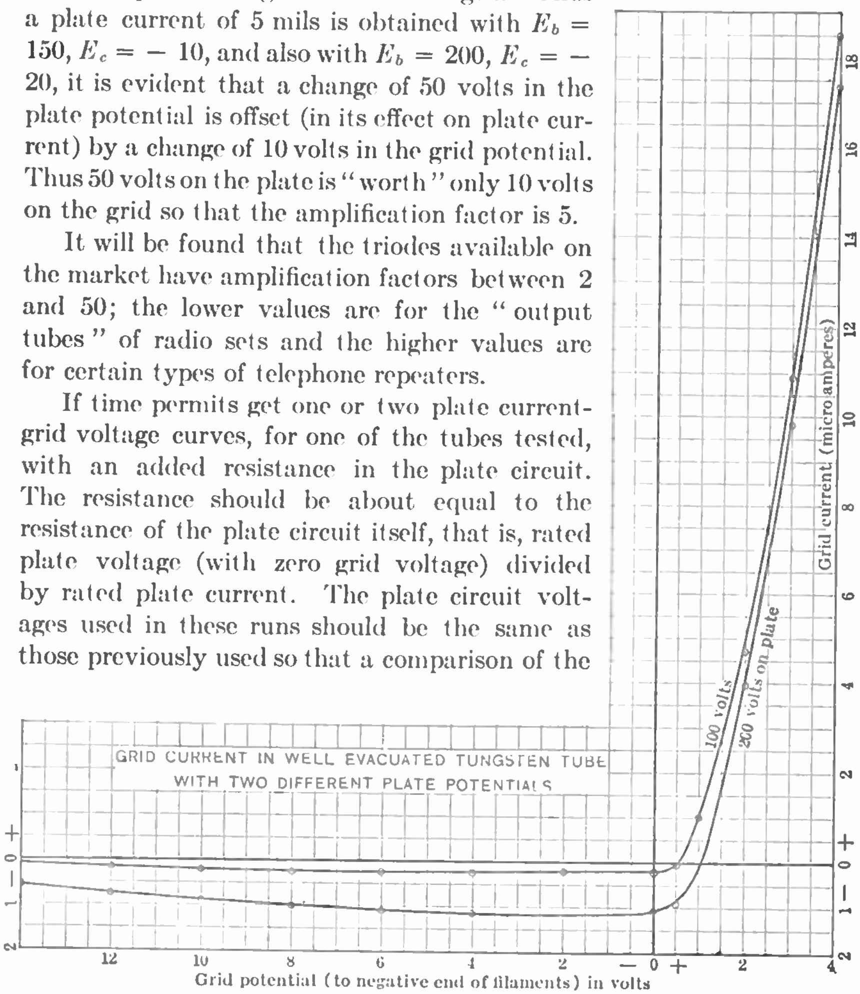


Fig. 104.

EXPERIMENT 21

Object.—Study of characteristic curves of tetrodes and pentodes.

Analysis.—A very serious trouble is caused in a triode, operating as a radio frequency amplifier, by the electrostatic capacity between the grid and the plate. Although the capacity is only a few micro-microfarads, its action in coupling the output and input circuits of the amplifier stages is generally sufficient to sustain self-excited oscillations in these circuits. And a triode in which such self-excited oscillations exist is practically useless as an amplifier.

Also the effective capacity of the input circuit of a triode is very much greater than the geometrical capacity if the tube is arranged to amplify. The effective input capacity may be several times as large as the geometrical capacity—the value measured in Exp. 12.

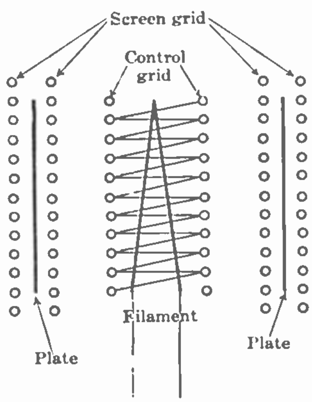


Fig. 105.

Both these effects may be reduced to practically zero by interposing another grid-like structure between the grid and plate, that is, putting in another grid. In the type of tube being discussed here this structure takes the form shown in Fig. 105; this is called the screen grid tube. The extra grid is made up of two helical grids, one just inside, and one just outside, the plate, these two being connected together as one grid. It will be seen that this structure completely surrounds the plate with the screen grid.

The scheme of connections used with such a screen grid tube is shown in Fig. 106; the screen grid is connected directly to a suitable point on the *B* battery with no coil or resistance in between. As the negative end of the *B* battery is grounded it is evident that the screen grid potential cannot change from that steady constant value given by that portion of the *B* battery between it and ground.

Hence no matter how much the plate may fluctuate in voltage, setting up a fluctuating electric field, this fluctuating field cannot "reach through" the screen grid to the other grid (the control grid) and so cannot transfer energy from the plate circuit to the input circuit. That is, this constant potential grid (the screen grid) entirely eliminates the capacitive coupling between the plate circuit and the input circuit.

Furthermore, although the introduction of this screen grid does increase the geometrical capacity of the tube over the value it would have if the screen grid were not used, this value is not augmented by anything that happens in the plate circuit, so that the effective capacity of the input circuit is no more than the geometrical capacity. Thus the interposition of this screen grid between plate and control grid overcomes both the defects mentioned above regarding the triode.

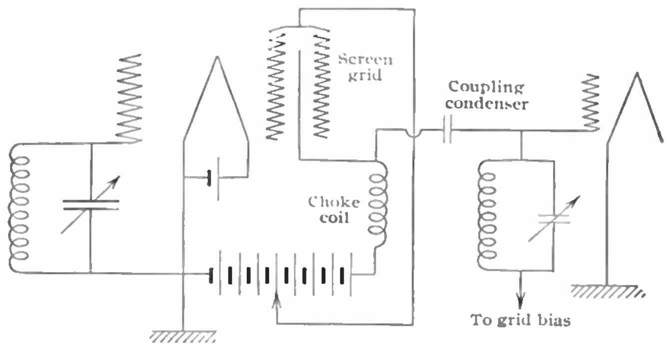


FIG. 106.

In carrying out the tests on the screen grid tube the student will encounter one of the most interesting of the phenomena which occur in vacuum tubes, namely, *secondary emission.* If an electron strikes an electrode with sufficiently high velocity it may " splash " out from the electrode several other electrons; the higher the speed of the impacting electron the more electrons it can splash out. These splashed-out electrons are called secondary emission.

Now ordinarily these electrons fall right back into the electrode from which they were knocked out, so we have no evidence of their ejection from the electrode. However, if there is another electrode close by, with a higher positive potential than the electrode in question, these splashed-out electrons may leave this electrode and travel over to the higher-potential one. And there may be more of these splashed-out electrons than there are doing the splashing. It may be that there are more electrons leaving the electrode than arriving, that is, the net flow of electrons is away from this positively charged electrode, and the electrode is cold! This phenomenon results in " negative plate current " an effect which seems anomalous until this phenomenon of secondary emission has been studied.

Let us suppose the control grid absent temporarily (as it plays no part in the phenomenon) and arrange the screen grid and plate as in Fig. 107. The screen grid G is held at a constant positive potential while that of the plate is continuously varied from zero up. The form of the actual plate current is complex, being the net result of two opposing actions. In Fig. 108 there is shown the form of plate current for voltages from zero to a value about equal to that of the screen grid: this is the curve $OEFGH$. This is the only curve that can be measured, but its form can be understood by the use of several other curves, constructed from theoretical reasoning. Curve OA shows the electron current to P due to emission from the filament; curve OB shows the amount of secondary emission from P, due to the electrons of current OA; curve C shows the fractional part of the secondary emission which is attracted to G; curve OD shows the electron current away from P due to secondary emission; and curve $OEFGH$ shows the actual plate current, which is the result of these component effects. The values of plate voltage assumed in plotting Fig. 108 are too low to make the tube of any service in the ordinary amplifying circuit; in Fig. 109 are shown the complete forms of plate and screen grid currents, as plate voltage is varied from zero to a value considerably higher than that of the screen grid.

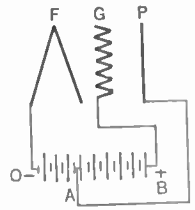


FIG. 107.

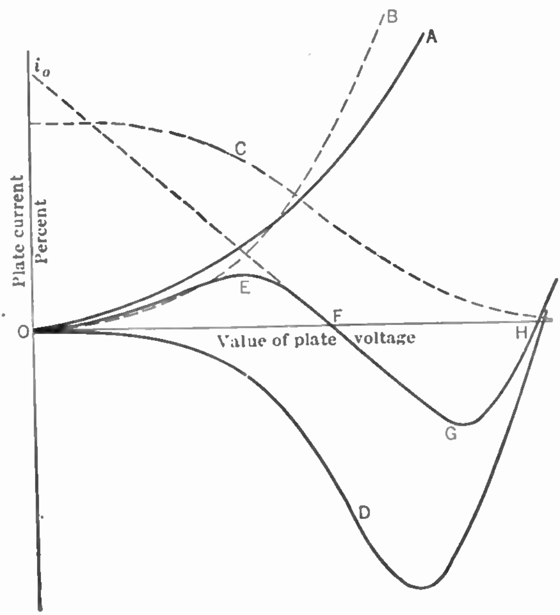


FIG. 108.

The first part of the plate current curve is the same slope as that of Fig. 108, but this is not the portion of the characteristic ordinarily used. In Fig. 109 it is seen that for voltage greater than OA the plate current increases uniformly with plate voltage, and in this region the screen grid current is reasonably low; this is the operating region of the screen grid tube.

In getting these curves the control grid potential is held fixed; the slope of the plate current curve therefore gives the a.c. resistance of the plate circuit of the tube. The plate voltage is relatively impotent in controlling plate current because of the screen grid interposed between the plate and the source of electrons. This makes the a.c. resistance of the plate circuit, $\Delta E_P/\Delta I_P$, very high, so much so that it is difficult to build a circuit which is suited to the needs of such a tube.

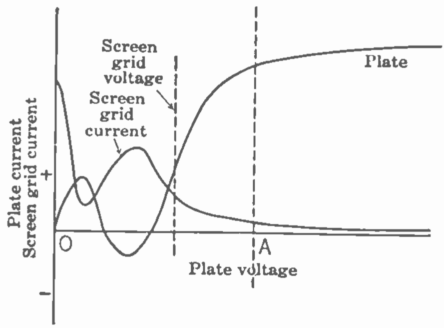


FIG. 109.

If the control grid potential is now varied, assisting and opposing the space charge, the currents to both screen grid and plate change greatly, keeping approximately the same division of total current between them. In other words, the screen grid in no way hinders the control grid from affecting the plate current, but does very greatly detract from the plate voltage control over plate current.

Actually a change of 1 volt in control grid potential may change the plate current as much as does a change of 200 volts on the plate; this then would give an amplification factor of 200. In actual tubes the amplification factor, defined as $\frac{\Delta I_P}{\Delta E_g} \Big/ \frac{\Delta I_P}{\Delta E_P}$, may be much more than that.

In general, there are many more electrons evaporated from the filament of a receiving tube than are required for the plate current. If with a given plate voltage, some arrangement results in a large increase in plate current, there is a considerable advantage gained because the plate circuit resistance is cut down.

Of course the control grid could be held at positive potential, and this will very materially increase the plate current for a given plate voltage, as evidenced by the results of Exp. 20. But such an expedient is not available because, for reasons which will appear in later experiments, the control grid, connected to the input circuit, must practically always be held at such a potential that it draws no current; this means that the control grid must be held at negative potentials.

However, it is possible to put in an extra grid, surrounding the filament and close to it, which is held permanently at some such positive potential that the space charge is partially neutralized, thus increasing the supply of electrons available for plate current. A grid so used is called a space charge grid, and, when used in a screen grid tube, evidently results in the pentode, or five-electrode tube. Surrounding the filament is the space charge grid; around this is the control grid; next comes the screen grid surrounding the plate, and finally the plate. In case a heater type of tube is used there are required two terminals for the heater, one for the cathode, one for each of the three grids, and one for the plate or a total of seven terminals. The control grid terminal generally comes through the top of the bulb (in American practice), the socket has five terminals arranged as usual for heater type tubes, and one terminal comes through the side of the tube base.

Get characteristic curves of the screen grid tube for normal plate voltage and screen grid voltage (as well as normal filament conditions), reading currents to screen grid, control grid, and plate as control grid potential is varied through suitable limits.

Get similar curves for normal plate volts and decreased screen grid voltage, and for normal screen grid volts and decreased plate voltage.

With screen grid at rated voltage, control grid constant at about 2 volts negative, get curves of screen grid current and plate current as plate voltage is varied from zero to 25 per cent above its rated value.

Get similar curves with the control grid held at zero potential.

With a pentode get curves of currents in plate, screen grid, control grid, and space charge grid, as control grid is varied with plate and screen grid at normal potentials and space charge grid at zero potential. Get a similar set with the space charge grid at some suitable positive potential as recommended by the manufacturer.

EXPERIMENT 22

Object.—Study of the vacuum tube voltmeter and effect of frequency upon its calibration.

Analysis.—In the radio laboratory the ordinary a.c. voltmeter is of very little use, primarily on account of the large current required for its operation, and secondarily because of the very great error in its indications when the frequency is high. The hot wire type is free from this frequency error, but the power required for its operation is even greater than that required for the ordinary iron vane type.

Of course, as was pointed out in Exp. 14, the high-resistance thermocouple ammeter may well be used as a voltmeter; this type is free from frequency error and uses but a small fraction of the power required by the hot wire ammeter, but, even so, its power demands are so great that its use in many radio circuits result in errors of 50 per cent or more.

Let us consider that the most sensitive thermocouple voltmeter has a resistance of 1000 ohms, for a 1-volt range. If now we use this voltmeter to measure the voltage across the ordinary radio frequency tuning condenser, the potential difference will be about half as much when the voltmeter is connected as it was before the voltmeter was connected. Thus the value of voltage indicated by the meter is only one-half as much as it should be!

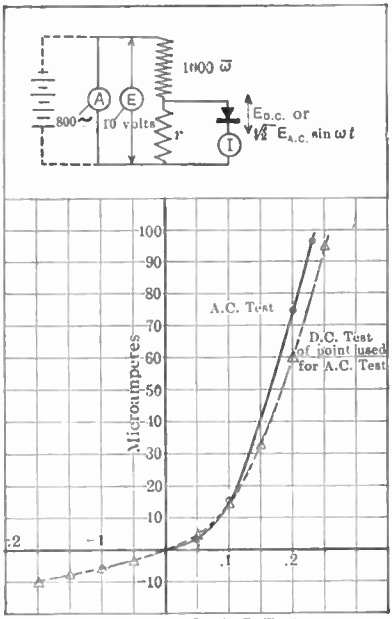
From considerations of this kind it appears that the resistance of a proper voltmeter for the radio laboratory should be of the order of 1,000,000 ohms. But such a meter, with a 1-volt scale, would draw only 1 microwatt, and such an infinitesimal power is quite evidently insufficient to actuate a meter movement.

However by using a triode, suitably connected, it is possible to make this amount of power control the power output of a *B* battery, and thus give readable deflections on an ordinary continuous current meter. This is the principle of the triode voltmeter.

There are several types available, but they all operate on the idea of impressing the voltage to be measured between the grid and cathode of a triode, and measuring the effect of this voltage on the plate circuit of the triode.

Unless the grid is allowed to become positive with respect to the electron-emitting cathode, the a.c. resistance between grid and cathode is of the order of 1,000,000 ohms. As the capacity between grid and cathode is about 5 micro-microfarads, its reactance at one megacycle cycle is about 30,000 ohms, hence the impedance of the input circuit is about 30,000 ohms with a characteristic angle of about 88°. Such a voltmeter will, therefore, draw but little power from the radio circuit being measured. It will have an appreciable effect on the tuning of a sharply resonant circuit, but this effect can be compensated by a suitable readjustment of the circuit itself.

It may occur to the student that a crystal contact (such as zincite-chalcopyrite) in series with a sensitive portable galvanometer might be used as a voltmeter. In Fig. 110 is shown the relation between rectified current and impressed alternating voltage for a sensitive crystal contact. If such a contact could be kept in permanent adjustment the arrangement would be reasonably good as a sensitive voltmeter, but any excess voltage, or slight jar, completely spoils the point and so requires that another point be found and calibrated.



Scale of Volts for D.C. Test
Effective values for A.C. Test

FIG. 110.

The copper-copper oxide rectifier offers possibilities, but as arranged in a voltmeter its action is not independent of frequency. It is found to have an error as great as 10 per cent for frequencies of only 10 kc. Its alternating current resistance is about 2000 ohms per volt, but its impedance at high frequencies drops to a very small fraction of this amount because of the capacity by-pass existing in the meter circuit.

Thus we return to the triode as the most suitable device for the voltmeter. In Fig. 111 are shown the changes which occur in the average value of the plate current in an ordinary amplifying triode, as an alternating voltage is impressed between grid and cathode. (In the same figure is shown the corresponding effect obtained from a sensitive rectifying crystal contact.)

When the alternating voltage is applied directly between grid and cathode the average value of the plate current *increases* as the impressed voltage increases. If a grid condenser and leak are used the average value of plate current *decreases* as the impressed voltage is increased; this decrease is about three times as great as the increase, for a given voltage. It would thus appear that the triode with grid condenser and leak would make the most sensitive voltmeter, and this is so; but this scheme results in a frequency error which does not occur when using no grid condenser; the one type is therefore about as valuable as the other, its advantages compensate for its disadvantages.

By inspection of Fig. 111 it will be evident that unless some special arrangement is used it is difficult to read accurately voltages less than about 1 volt. A triode such as used in getting the curves of Fig. 111 draws a normal plate current of about 1 milliampere. If, now 0.5 volt (alternating) is impressed between grid and cathode the current will increase 10 *microamperes.* Thus the plate ammeter would indicate an increase of only 1.0 per cent—a scarcely perceptible change!

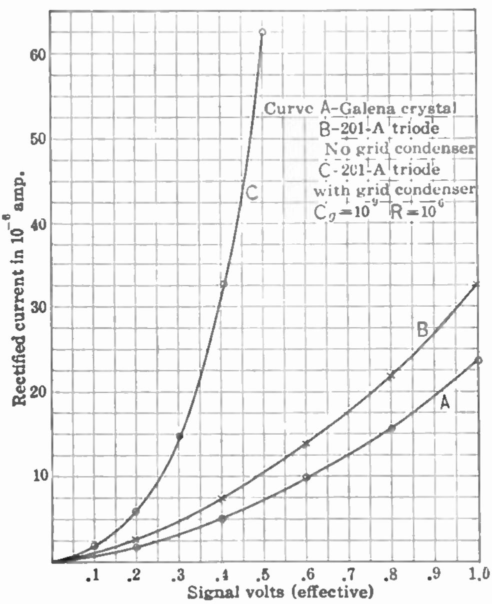


FIG. 111.

An obvious improvement requires that the normal current (the 1 milliampere) be shunted around the meter in some way, so that this meter reads only the *change in plate current.* Such a scheme is shown in Fig. 112. The normal plate current of the tube, caused by battery *B*, is just compensated (in the ammeter) by another current in the opposite direction through the ammeter produced by battery *C* through the resistance *R* and the ammeter in series.

Suppose the ammeter *A* has a resistance of 10 ohms and that battery *B* is sufficient to cause a plate current of 1 milliampere. If now *R* has 990 ohms resistance and battery *C* has 1 volt e.m.f., a current of 1 milliampere will flow through *R* and *A* in series and will neutralize the 1 milliampere of plate current in *A*. In other words, the 1 milliampere of plate current will flow through *R* instead of through *A*.

If now the plate current increases, this increase will divide between R and A in the inverse ratio of their resistances. This means that practically all of the increase in plate current will flow through A. Thus A may be of sufficient range to measure the increase in plate current only and thus make possible much greater accuracy than would be obtained without this compensating circuit.

If the voltage of battery C cannot be accurately adjusted to bring about complete compensation of the normal plate current, the resistance R can be made adjustable to a small extent, or R may be connected to the slider of a potentiometer connected to battery C. Of course, any considerable change in the value of R would result in some change in the calibration of the meter.

With such a compensating scheme as that being discussed it is evident that any change in filament current, plate voltage, etc., will very

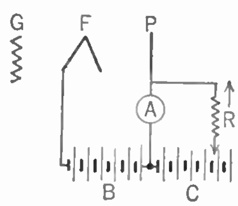


Fig. 112.

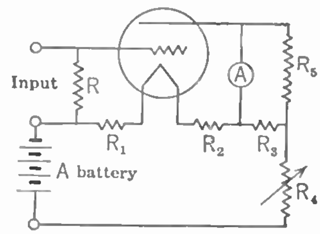


Fig. 113.

seriously affect the zero adjustment. An arrangement which acts to stabilize the zero setting to some extent in shown in Fig. 113. The input posts of the voltmeter are connected together by resistance R, a grid leak of about 5 megohms. This prevents a " free " grid even if the circuit being measured has a condenser between the two points connected to the voltmeter. The resistance R_1 serves to give a proper bias to the grid. The R_1I drop should be equal to the peak value of the largest voltage to be measured.

The plate voltage is obtained by the drop through R_2. The drop through R_3 gives the compensating voltage; the R_3I drop sends a current through R_5 and A which is equal to the normal plate current of the triode, to make ammeter A read zero. The resistance of R_5 should be many times the resistance of the ammeter A.

The variable resistance R_4 should be sufficient, in connection with R_1, R_2, R_3, and the filament resistance to limit the current from battery A to the normal filament current of the triode. This scheme uses only one battery; if the voltage of this battery falls both plate voltage and bias voltage decrease proportionally, tending to hold the zero setting of ammeter A constant. After the proper values of R_1, R_2, R_3, and R_5 have been selected, the zero setting of ammeter A is always obtained by varying R_4 only. Its calibration is the same for all frequencies up to some megacycles; it is adjusted ready for use by merely changing R_4 to make A read zero.

In another arrangement of triode voltmeter the compensation for zero adjustment is obtained by the sliding contact of a potentiometer P connected across the A battery, as in Fig. 114. A condenser C_1 serves to by-pass the high-frequency fluctuations of plate current around the ammeter A and the B battery.

In what looks like another type of triode voltmeter a Wheatstone bridge circuit is set up with the plate resistance of the triode as one arm

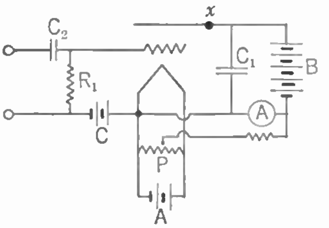


FIG. 114.

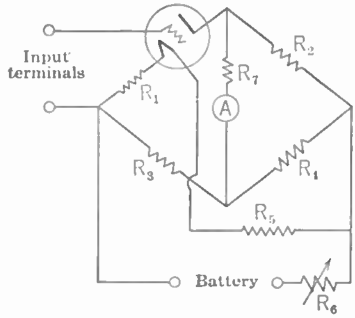


FIG. 115.

of the bridge, as shown in Fig. 115. One battery serves to give plate voltage, filament current, and grid bias. The ammeter A is brought to its proper zero reading by variation of R_6 only; all other resistances are left fixed after the circuit is once properly set up. A triode of low filament current is used, and one $22\frac{1}{2}$ volt B battery serves for plate, grid, and filament power. This type of meter is said to maintain its calibration for 1000 hours and to have a frequency error of only 3 per cent up to 300 kc. That is, if calibrated on a 60-cycle line its readings may be reasonably well relied upon, for frequencies nearly as high as the broadcast frequencies.

If a triode voltmeter is to maintain its high input resistance the grid must never swing positive; this means that the impressed voltage must never exceed the amount of grid bias. For convenience, low plate voltages are used, with triode voltmeters, and hence rather small bias voltages may be used, otherwise there would be no plate current. These small biasing voltages mean that only low values of alternating voltage can be measured; the ordinary one has a full scale reading of about 3 volts.

To extend the range a voltage divider of high resistance or a series connection of condensers may be used, the triode voltmeter being connected across one of them. In Fig. 116 *a* shows a potentiometer for a multiplier of 10. The impedance of the triode voltmeter must be high compared to 10^5 ohms if this arrangement is to give accurate results. In Fig. 116*b* two condensers are connected in series to the voltage being measured. By having the condenser C_2 with about 9 times the capacity of condenser C_1 a multiplier of 10 is obtained here also. The reactance of C_2 must be small compared to the impedance of the triode voltmeter.

Of course it must be realized that the impedance of the circuit making up the potential divider must be high compared to the impedance of the circuit to which it is attached, otherwise the potential difference

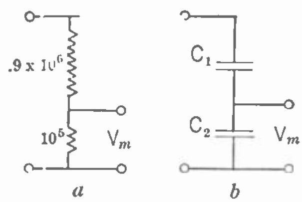


FIG. 116.

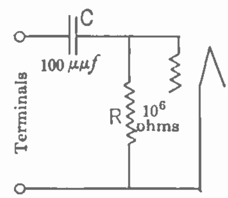


FIG. 117.

which is to be measured will be appreciably altered by the connection of it to the circuit.

A simple analysis shows that the triode voltmeter which uses no condenser in series with the grid should have the same calibration at a megacycle as it has at 100 cycles (or any other frequency); the impedance of the external plate circuit should of course be independent of frequency if this is to be true.

Any circuit arrangement which uses a series grid condenser, and grid leak, can be expected to have a calibration varying with the frequency. The triode can respond only to that voltage which is impressed on its grid, and only a fraction of the voltage impressed on the voltmeter terminals appears at the grid. This idea is brought out in Fig. 117; a grid leak of 1 megohm is used with a condenser of 100 micro-microfarads. For simplicity's sake we neglect the capacity and leakage between the grid and filament of the tube itself.

At one megacycle the reactance of the grid condenser is about 1700 ohms, so that practically all of the voltage impressed on the terminals will appear across the megohm, that is, between the grid and filament of the triode. However, if the frequency of the voltage being measured is 60 cycles the reactance of the grid condenser is about 25 megohms so that only about 4 per cent of the voltage impressed on the terminals appears across the grid-filament circuit of the triode. Its indications therefore will be only a small fraction of what it should be.

As the rectified current of the plate circuit varies with the square of the alternating voltage impressed on the grid, the calibration curve of the triode voltmeter must be of the parabolic type, having a greatly condensed calibration on the lower part of the scale.

The form of calibration can be somewhat altered if a resistance is inserted in the external plate circuit; this scheme of course makes the voltmeter less sensitive in addition to altering the form of its calibration curve.

With an ordinary triode and suitable variable resistances and ammeters make up a triode voltmeter and calibrate it on low frequency, intermediate frequency, and high frequency. Do this with no condenser in series with the grid, as well as with a suitable grid condenser and leak. For measuring the voltage impressed on the triode voltmeter use a high-resistance thermocouple, which has been calibrated as a voltmeter. This can, of course, be done on continuous current, as a properly constructed thermocouple has a negligible frequency error (at least for frequencies as high as 10^6 cycles).

Try the effect of using different capacity condensers for the grid condenser of the voltmeter.

Try the effect, or both sensitivity and form of calibration curve, of inserting in the plate circuit of the triode a resistance about equal to the resistance of the plate circuit of the triode. This should be inserted directly in series with the plate, as at X in Fig. 114.

After this work with the triode connected as a voltmeter, calibrate on both high and low frequency (say 1000 kc. and 1 kc.) one or two of the commercial thermionic voltmeters in the laboratory.

To show the action of the modern "power" detector find the relation between plate current and input voltage when the triode has a 150 volt plate battery, grid condenser, grid leak of about 5 megohms, in series with enough biasing battery to give a plate current of about .001 ampere, with no signal voltage impressed. In this run the input voltage is to be increased (in about ten steps) to about 20 volts. The plate current (as read on the plate circuit c.c. ammeter) should increase linearly with the signal voltage.

EXPERIMENT 23

Object.—Study of the triode arranged as a detector, both with and without a grid condenser. Comparison of triode with crystal or other form of detector.

Analysis.—A detector is a device which will give a unidirectional current when acted upon by an alternating voltage. The amount of unidirectional current must correspond to the amplitude of the alternating voltage, so that if a high-frequency voltage of varying amplitude is impressed on the detector there results a unidirectional current the amplitude of which corresponds to the amplitude of the alternating voltage. The sensitivity of a detector depends upon the amount of rectified current it yields for a given amplitude of alternating voltage.

The amount of rectified current the device yields should be proportional to the amplitude of the impressed voltage, but the ordinary detector gives a rectified current which varies approximately with the square of the impressed voltage, when this is small, as was brought out by the result of Exp. 22. This makes it a very ineffective device when the voltage is low, just the case where the efficiency of the detector should be a maximum. This is a defect common to practically all forms of detectors.

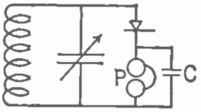


FIG. 118.

In addition to being sensitive a detector should be rugged, and not easily disturbed from its proper adjustment. The crystal detector is almost as sensitive as the triode, but of course the crystal contact is easily disturbed, or its rectifying property spoiled by a strong signal burning it out, so that from the standpoint of reliability the triode is far superior to the crystal. However, the battery power required for its operation makes the triode more costly to maintain, so that for small, cheap receiving sets the crystal detector is still used.

The detector is generally connected in shunt with the variable condenser of the tuning circuit, the detector being directly in series with the telephone receiver in the case of the crystal. This brings about another requirement of the detector, as can be seen by analysis of Fig. 118. The phone is generally shunted with a by-pass condenser C, the reactance of which is very small for the high-frequency current in the tuned circuit. Thus, in so far as the high-frequency currents are concerned, the shunt path consists only of the a.c. resistance of the detector.

The a.c. resistance of the rectifier must therefore be high, or else the selectivity of the tuned circuit will be spoiled. The effect of the shunt resistance is most easily determined by changing it to an equivalent series resistance by the relation $R_{\text{series}} = X^2/R_{\text{shunt}}$, in which X is the reactance of either the coil or condenser in the tuned circuit. A common value of such reactance at broadcast frequencies is about 800 ohms; if therefore the a.c. resistance of the rectifier is 10,000 ohms its equivalent series resistance is 64 ohms. Now if the power factor of the coil is 2 per cent (an ordinary value) its resistance is 16 ohms. It follows therefore that the effect of shunting the rectifier circuit around the tuning condenser is to increase the equivalent resistance of the tuning circuit from 16 to 80 ohms, that is, increase the width of its resonance curve five times!

It is thus quite apparent that the effect of the detector circuit is to very materially decrease the selectivity, and that unless the a.c. resistance of the detector is of the order of 100,000 ohms the selectivity of the circuit is reduced to a small fraction of the value it has without the detector connected.

In case the triode is used as detector the grid-filament resistance is connected across the tuned circuit; this resistance is practically never as low as 10,000 ohms (a quite common value for a crystal) but is generally in excess of 100,000 ohms. It will thus be found that the selectivity of a circuit using a triode detector is always better than one using a crystal.

As was found in Exp. 22, the rectifying quality of a triode is considerably better when a grid condenser and leak are used in the input circuit, than when the a.c. signal is applied directly between grid and filament, and such is nearly always the arrangement of the triode when used as a detector. The grid leak generally is returned to the positive filament terminal, if there is such; even though the input resistance is somewhat reduced by thus making the grid slightly positive, the rectifying quality is so much better (than with a negative grid) that the bad effect of the lowered resistance is more than compensated.

The triode used as detector has usually employed a plate voltage of 45 volts or less, and the amplitude of signal which could be impressed on the input circuit was only a few volts. If a greater signal was impressed a disagreeable distortion of the quality of the signal in the telephones resulted. The tendency in modern broadcast receiver design, however, is to use a much higher voltage on the plate (up to 150 volts) and, with proper values of grid condenser and leak, to "detect" signals as high as 20 volts or more.

In this experiment a damped wave signal is to be generated by a buzzer wave generator, and this signal is to be picked up by a loosely coupled tuned circuit, across which either the triode or crystal rectifier can be used as detector. As the entire wiring arrangement is to be the same except the detector, the relative signal strengths heard in the telephone receiver serve to indicate the relative efficiencies of the detectors. Further, the sharpness of tuning of the receiving circuit when using first one and then the other detector serves to show the relative damping effects produced by the two.

This is the first experiment of the course in which the student is asked to use his ears as a quantitative measurer of sound, and he will soon be convinced how poorly the ear is adapted to this task. As a qualitative meter of sound the ear is most remarkable, giving a reliable response for sound power as low as about 10^{-12} watt, but when it is required that the experimenter use his ear to compare the relative loudness of two sounds the ear is almost useless until after much practice. Listening to first one signal and then to another, one experimenter will judge that one is twice as loud as the other, while another listener will say that the one is perhaps a hundred times as loud as the other. If the strength of a signal is changed from a certain value to one five times as strong, a person unskilled in sound measurement may guess that it has been increased perhaps 10 per cent!

Although the ear is so poor in judging how loud one sound is, compared to another, it is comparatively easy to judge when two sounds have the same intensity. By arranging the phones to be connected alternately to one circuit and then to the other, arrangements being at hand for adjusting the strength of one of them, equality of signal strength can be obtained by ear within about 10 per cent.

For adjusting the strength of a signal by a definite amount, a type of telephone shunt styled an " audibility meter " has been much used. It consists of a double resistance unit, with two moving contacts operated by the same lever; the scheme is indicated conventionally in Fig. 119. The two resistances A and B are connected together by the bar joining the sliding contacts C and D. This bar also constitutes one of the terminals for connecting the telephones. With the contacts at the extreme left the signal on the phones is a maximum, and as they move to the right the phones are connected across a decreasing part of resistance B; at the same time more of resistance A is connected in the input circuit.

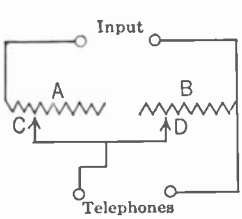


FIG. 119.

By properly proportioning the relative resistance in the successive resistance steps to that of the phones, it is possible for the device to show a constant resistance across the input terminals, for all positions of the variable contacts. The successive resistance steps are calibrated in terms of audibility; the contacts are moved until the phones give a just audible signal, and the setting gives the "audibility" the signal would have if the phones were unshunted.

Even when using this convenient device, the measurement of the audibility of a signal is a most inaccurate determination, because as the test proceeds and the ear becomes accustomed to the weak signals, the audibility of a given strength of signal, as determined by the meter setting, will increase many times. This change in the sensitivity of the ear is of no importance, however, when two signals are being compared for *relative* audibility, as the change in the ear affects both alike.

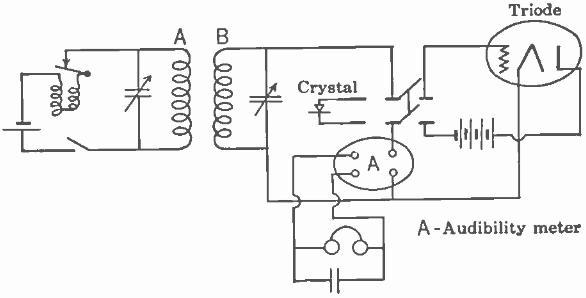


FIG. 120.

With the apparatus set up about as in Fig. 120, and loose coupling between the two tuned circuits, the phone is switched back and forth from triode to crystal, and the latter is adjusted to give a point having as good rectification and as small a damping effect as can easily be found. During this adjustment the triode serves merely as a standard in comparing the various points tried on the crystal, until a good one is found.

Then, being careful not to spoil the crystal adjustment, the various connections of the triode are compared to the crystal as standard, the audibility being adjusted to give the same convenient signal strength for both detectors. The ratio of the two audibilities gives a measure of the relative efficiencies of the crystal and triode as detectors, and the width of the setting of the variable tuning condenser over which the signal can be heard gives a qualitative measure of their damping effects.

The wiring of both circuits, sending and receiving, must be compact and orderly, and so arranged that no appreciable signal is induced in the receiving circuit except where intended, between the two coils A and B. The wiring of the triode detector is necessarily more extensive than that of the crystal, and so unless care is exercised there will be extra voltages induced in the triode circuit. These extra voltages may assist, or may oppose, the voltage induced in the coil of the tuned receiving circuit, thus making the comparison of results meaningless. The triode wiring should be neatly carried out, compactly arranged, and as distant from the wave generator as the compact wiring arrangement permits.

To test how much of this undesired induction is taking place, the two coils *A* and *B* can be placed with their axes at right angles in such relative positions that their coupling is essentially zero—the remaining signal must be due to the extraneous voltages. The wiring should be arranged so that this residual signal is small compared to that picked up by the coupling of the two coils, as these are normally arranged for carrying out the test.

When the circuits have been satisfactorily arranged, compare the triode to the crystal, for various plate voltages and grid bias, with no grid condenser and leak. Then using various-size condensers and grid leaks, and plate voltages, with leak connected to negative, as well as positive filament terminal, make sufficient comparison with the crystal to bring out the merits of the various adjustments.

If time permits, see whether the pitch of the received signal has any significance in this test; this change in pitch can be obtained by changing the tension on the buzzer armature. With a high-pitched signal it should be found that the optimum grid condenser has a smaller capacity than for a low-pitched signal, this on the assumption that grid leak and other conditions are not varied.

As an extra feature of this test arrange the telephones for potentiometer excitation from a constant voltage source of some kind, as in Fig. 121. As the resistance across which the phones are connected is increased, the current through the phones increases proportionately, and as the power supplied to the phones varies with the square of the current through them, and the efficiency of the phones is practically independent of current, the sound output of the phones should vary as the square of the resistance across which the phones are connected. While one student varies the setting of the potentiometer the other is to try to estimate the relative signal strengths and then compare his estimate with the known ratio as obtained from the potentiometer setting.

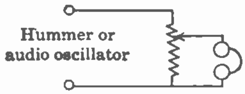


Fig. 121.

EXPERIMENT 24

Object.—To measure the amplification factor and plate resistance of a triode and their dependence upon tube construction, plate voltage, filament current, and grid bias. Mutual conductance. Available voltage amplification.

Analysis.—The great utility of the triode depends primarily upon the fact that the flow of electrons from cathode to plate can be easily controlled by the potential of the grid, and that the control of the current in the plate circuit can be brought about with a practically negligible power expenditure in the grid circuit.

The plate current can thus be controlled by varying either the plate voltage (as was shown in Exp. 19) or grid voltage. If the plate current is increased an amount ΔI_p by an increase in plate voltage equal to ΔE_p (grid potential remaining constant) and is then brought back to its original value by a decrease in grid potential equal to ΔE_g it will be appreciated that, in so far as plate current is concerned, a change in grid potential of ΔE_g has the same effect as a change in plate potential of ΔE_p. As triodes are ordinarily constructed, ΔE_p is much greater than ΔE_g.

The relative effectiveness of grid and plate potentials, in controlling the plate current, is indicated in radio literature by the symbol μ. It is defined by the relation

$$\mu = \Delta E_p/\Delta E_g (I_p \text{ constant}) \quad . \quad . \quad . \quad . \quad . \quad . \quad (53)$$

For the tubes available on the market the factor μ varies from about 3 to 50. The finer the mesh of the grid and the closer it is to the plate, the greater is the value of μ. It is difficult to derive an exact formula for μ from the geometrical constants of the triode unless this is assumed of some simple geometrical form. On the assumption that cathode, grid, and plate are all infinite plane surfaces, it has been shown that

$$\mu = 2\pi an/\log_e \frac{1}{2\pi rn} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (54)$$

in which a = distance between grid and plate, in centimeters
n = number of grid wires per centimeter
r = radius of grid wires

Even though the form of the electrodes of actual triodes is far from that assumed in deriving this formula, the values of μ obtained experimentally and from the formula do agree within about 10 per cent.

The resistance of the plate circuit of the triode, that is, the resistance between plate and cathode, is of a peculiar kind and serves well to bring to the student's attention the fundamental nature of all electrical resistance. If we define resistance from the fundamental concept, $R = \text{Heat}/I^2$, the question at once arises, where is the heat generated as the electrons flow from cathode to plate?

Starting from the cathode surface with a relatively low velocity, the electrons accelerate in the inter-electrode space until they have velocity measured in thousands of miles per second when they reach the surface of the plate. In this inter-electrode space no heat (heterogeneous molecular motion) is generated as there are no molecules present to participate in the electrons activity. (Of course if sufficient gas is present to interfere with the flow of electrons this statement is not true, but on the other hand, such a triode is not a true vacuum tube.) *The space between cathode and plate thus has no resistance.*

When the electrons collide with the plate, however, they must come to rest, and here they give up their kinetic energy to the molecules of the plate; the plate becomes hot as a result of this action, and it is this amount of heat which, divided by the square of the plate current, gives the resistance of the cathode-plate circuit. This evidently is the nature of all resistance. The electrons, progressing through the material of an ordinary wire conductor, collide with its molecules and give up their kinetic energy to the molecules; the energy thus gained by the molecules appears as heat in the conductor.

The amount of heat generated at the surface of the plate is equal to the plate current multiplied by the voltage between plate and cathode. This product really gives the potential energy (per second) lost by the electrons as they " fall " from the cathode to the plate, but in an action of this kind the potential energy lost is equal to the kinetic energy gained. Thus the kinetic energy given up to the plate by the rapidly moving electrons being suddenly brought to rest is equal to the potential energy they gained in falling from cathode to plate, or E_pI_p. The resistance then is equal to E_p/I_p. The value of resistance so calculated is the continuous current resistance, designated by R_{op}. In the average tube, it varies inversely with the voltage on the plate.

In general, the plate current of a triode is a fluctuating one; this fluctuating current is to be considered as an alternating current superimposed on a constant continuous current. Experimentally the fluctuating unidirectional plate current is actually split up into these two components, as in Fig. 122. Here the alternating current $I_{a.c.}$ flows through the condenser by-pass around the choke coil, which carries the constant current $I_{c.c.}$.

The resistance of the plate circuit (plate to cathode) of the triode for the alternating current is defined as

$$R_p = \Delta E_p/\Delta I_p \quad \quad (55)$$

As the plate current is not directly proportional to the plate voltage this resistance is not the same as R_{op}; it is about one-half as much for the ordinary triode.

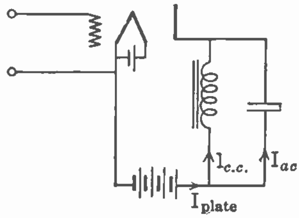


FIG. 122.

In determining the two resistances, thus far discussed, it has been assumed that the grid potential has been held at some fixed potential, generally taken as zero. Zero potential is taken as that of the cathode, for heater type tubes, as the potential of the negative end of the filament in a c.c. filament tube, and as the potential of the mid-point of the filament in the case of a filament heated by alternating current.

If the circuit of the triode has been so arranged that the external plate and grid circuits are coupled together, it may well be that when the plate voltage varies, thus varying the plate current, the grid potential is simultaneously varied by induced voltage of some sort. Then the change in plate current will be due to both ΔE_p and ΔE_g, and it will quite evidently be different than if ΔE_p were acting alone.

If then we write $R'_p = \Delta E_p/\Delta I_p$, remembering that ΔI_p is being controlled by ΔE_g as well as ΔE_p, it may well be that ΔE_g counteracts the effect of ΔE_p, so that ΔI_p is equal to zero. It then follows that R'_p is infinity! Or it may be that the effect of ΔE_g is to actually diminish I_p even though ΔE_p is positive. In such a case ΔI_p is negative when ΔE_p is positive, so that R'_p is negative! This negative value of R'_p is the fundamental requirement for a triode to generate alternating currents, which is one of its major roles.

In this experiment R_o is easily calculated from the readings of properly connected c.c meters, and R_p is measured by one of the balances obtained during the test. The measurement of R'_p is left for a subsequent experiment.

In Fig. 123 the potentiometer P serves to adjust the grid bias to any desired value; the plate voltage is varied by using more or less of the B battery. Two resistance boxes R_1 and R_2 are connected in series for excitation from the hummer, or other small a.c. generator. Grid, filament, and plate are connected to these resistance boxes as shown.

By this connection scheme it is evident that the alternating voltage impressed on the electrodes is opposite in phase; when the plate is going positive, with respect to the filament, the grid is going negative. With R_1 fixed at a small value, say 5 ohms, R_2 is varied until no note is heard in the phones; this means, of course, that no fluctuation is taking place in the plate current. And if no fluctuation is taking place in the plate current the ΔE_g and ΔE_p impressed by the hummer must be just neutralizing one another. But ΔE_p is given by IR_2 and ΔE_g by IR_1, so that we must have

$$\mu = R_2/R_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (56)$$

In deriving this relation it is assumed that all the current, I, delivered by the hummer goes through R_1 and R_2. This is practically always the case, as these will generally have values of less than 100

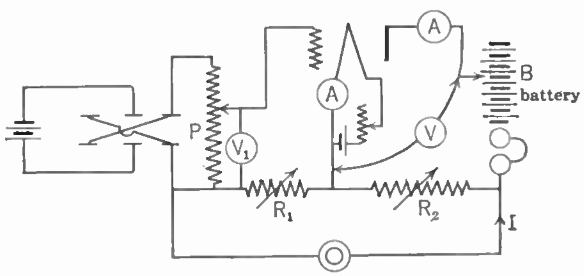


Fig. 123.

ohms, whereas the parallel paths (plate to filament and grid to filament) through the triode have many thousands of ohms.

The value of ΔE_g used in this test must be kept low if a good balance is to be obtained; the value of μ varies slightly for different values of grid voltage, so that if ΔE_g is made too large no value of R_2 can be found, which produces complete silence. For ordinary triodes ΔE_g should be less than 1 volt.

With the set up of Fig. 123, determine if, and how, μ varies as grid bias, plate voltage, and filament current are changed. Get sufficient points in each run to draw accurate curves showing the variation.

As ordinarily used in its role as amplifier, an impedance of some kind is inserted in the external plate circuit of the triode; the fluctuating plate current set up by variations in grid potential set up varying voltage across this impedance, and this varying voltage is greater than the voltage impressed on the grid. The ratio of this drop across the external impedance, to the input voltage impressed on the grid, gives the voltage amplification available for impressing on a succeeding triode or other device; it is called the voltage amplification of the triode.

The circuit arrangement which is convenient for measuring this voltage amplification is also suited for measuring the a.c. resistance of the plate circuit (internal) of the triode, so we will discuss the first. In Fig. 124 the essential parts of the test circuit are shown; filament battery, grid bias, etc., are omitted. The hummer impresses a voltage E_g between the filament and grid, and we know that, in so far as effect on plate current is concerned, this is equivalent to introducing a voltage equal to μE_g directly in the plate circuit.

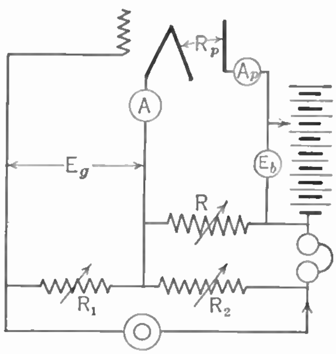


FIG. 124.

If in the circuit of Fig. 124 the phones give no signal, then all the alternating current flowing through the plate circuit of the triode is confined to the path made up of R and R_p in series.

Now let us suppose that R_2 is adjusted equal to R_1 and that R has been varied until there is no sound in the telephones. This balance requires that the alternating voltages across R_2 and R are equal.

Now the alternating current in the plate circuit must be given by $I_p = \mu E_g/(R + R_p.)$ The drop across the resistance R must be equal to I_pR or $\mu E_g \dfrac{R}{R + R_p}$. If now R_2 is equal to R_1 the drop across R_2 must be the same as that across R_1, that is, equal to E_g. So for the condition of balance we have

$$E_g = \mu E_g R/(R_p + R)$$

or

$$R_p = (\mu - 1)R. \quad \ldots \ldots \ldots \quad (57)$$

This simple relation is obtained on the assumption that $R_1 = R_2$, but, of course, a similar derivation can be used for any relation between R_1 and R_2 that permits a balance to be obtained. A brief consideration shows that the ratio R_2/R_1 must be less than μ if a balance is to be obtained.

With the circuit arrangement of Fig. 124 properly completed to read filament current, plate current, and plate voltage, and arrangements made for impressing and reading a variable grid bias, with R_2 adjusted equal to R_1, vary R until the phones are silent. The plate circuit resistance is then given by eq. (57)

Find how this plate resistance varies with plate voltage, grid bias, and filament current. For each setting read the voltage and continuous current in the plate circuit, so that R_{op} can be calculated for each setting.

In all these tests the voltage E_g must be kept low (only a fraction of one volt) if a sharp balance is to be obtained. The alternating current in the plate circuit will be by no means sinusoidal if a large grid voltage is used; the I_pR drop will thus not be sinusoidal. But the drop across R_2 against which I_pR is to be balanced is practically sinusoidal, hence a perfect balance cannot be obtained. Upper harmonic voltages will be left unbalanced no matter what value of R is used to obtain a balance.

The voltage amplification of the triode, for various values of external plate circuit resistance, can readily be obtained with the arrangement of Fig. 124. Various values of R are used, and for each R_2 is varied until a balance is obtained. Then the voltage amplification of the triode and its circuit is given by the ratio R_2/R_1, as brief consideration will show. The continuous value of the plate current is to be maintained constant by varying the voltage E_b.

Measure this voltage amplification for about ten values of R, in suitable steps from about one-quarter the value of R_p to a value about three times as large as R_p.

In a way this test is somewhat misleading, even though the values of voltage amplification obtained are correct. It will be found that, to maintain the plate current, as read on a c.c. meter, constant as R is varied, comparatively small changes in the plate circuit battery are required. This is because the resistance R is shunted by the path made up of R_2 and the phones in series, and with high values of R most of the plate current (the continuous current component) flows through this low-resistance path.

When the triode is actually used as a resistance coupled amplifier, no such shunt path exists, and the plate circuit voltage must be increased proportionally as R is increased. If this is not done the increase in R results in a decrease in plate current, and a subsequent increase in R_p. Hence the antipicated gain in voltage amplification will not be obtained.

The amplification in this arrangement is given by

$$\alpha = \mu \frac{R}{R + R_p} \quad \text{. . . . .} \quad (58)$$

and of course if R_p increases when R is increased, the gains in α will be unexpectedly small.

The term "mutual conductance" has obtained a firm footing in radio literature in spite of the fact that it is not a well-selected term. It is generally indicated by the symbol g_m, and is defined as that factor by which the alternating voltage impressed on the grid must be multiplied to give the alternating current set up in the plate circuit.

If I_p = magnitude of alternating current in plate circuit
E_g = alternating voltage impressed on grid
μ = amplification factor of triode
R_p = alternating current resistance of plate circuit

then

$$I_p = G_m E_g \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (59)$$

but

$$I_p = \mu E_g / R_p$$

so that

$$G_m = \mu / R_p. \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (60)$$

The factor G_m is generally given in "microamperes per volt" and varies in the commercial triodes from a few hundred to a few thousand.

Calculate the mutual conductance for the various tests carried out in this experiment.

Plot curves for the three tests outlined in this experiment, showing how the factor investigated varies with the other factors involved.

If time and apparatus permit, the variation of the voltage amplification, α, with a variable inductive reactance in the plate circuit can be investigated.

The bridge scheme of Fig. 124 can be used when the external plate circuit contains resistance only; a reactance in the plate circuit would not permit a balance to be obtained, as the drop across R_2 would not be in phase with that across the plate circuit impedance. If, however, a vacuum tube voltmeter of suitable voltage range is available the voltage amplification can be measured directly, without a bridge. The input voltage can be measured with a thermocouple, calibrated as a voltmeter, and the amplified voltage across the plate circuit reactance can be measured by the thermionic (vacuum tube) voltmeter. A condenser (with a high insulation resistance) should be used in series with this voltmeter so that the voltage drop due to the continuous current component of the plate current is not registered, in addition to the desired alternating voltage.

EXPERIMENT 25

Object.—Measurement of the capacity and resistance of the input circuit of a triode, under various conditions of plate circuit impedance.

Analysis.—The input capacity of a triode is directly in parallel with the tuning condenser of the input circuit if the ordinary radio frequency amplifier, so that its measurement and variation with circuit adjustments is of considerable importance. Furthermore, whereas the input circuit of the triode ordinarily has a positive resistance of many thousand ohms, it may have an entirely different value under certain circuit conditions; it may in fact be negative. As indicated in Fig. 125, the capacity of the input circuit is made up of that between the grid and filament, in parallel with that between grid and plate. Thus the charging current I must be large enough to charge both condensers.

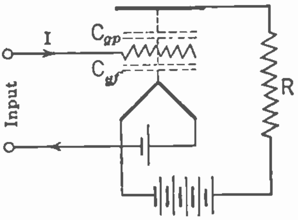


FIG. 125.

Now the voltage impressed on the condenser C_{gf} is evidently E_g, the input voltage, so the charging current required for C_{gf} is $\omega C_{gf} E_g$. Now the changing grid potential makes the plate potential rise and fall, owing to the fluctuations of plate current flowing through the plate circuit impedance R. When the impedance is zero the plate potential does not fluctuate at all, so that in so far as charging current is concerned, the plate acts the same as though it were part of the filament, and hence the charging current required for this part of the circuit is $\omega C_{gp} E_g$.

Now as the impedance of the plate circuit is increased the plate goes up and down in potential, with respect to the filament, by an amount equal to the potential drop in the impedance R. In the last experiment we showed that the voltage developed across R was equal to $\mu R/(R + R_p)$, which we called α. Now the plate potential fluctuates in phase opposition to that of the grid, that is, when the grid potential is increasing with respect to the filament, the plate potential is decreasing. This follows from a simple analysis of the circuit relations as shown below.

When the grid potential increases, the plate current increases, thus increasing the RI drop in the resistance R. But the plate potential is equal to the voltage of the B battery, minus the drop in resistance R, hence it follows that the plate potential is falling when the grid potential is rising.

From this it follows that if the voltage between plate and filament is equal to αE_g it must be equal to $(\alpha + 1)E_g$ between the plate and grid. Thus the condenser C_{gp} must be charged to a voltage $(\alpha + 1)E_g$ when the voltage E_g is impressed on the input circuit of the triode. The total charging current which must, therefore, be supplied from the input circuit is equal to $\omega C_{gf}E_g + \omega C_{gp}(\alpha + 1)E_g$, so that the apparent capacity of the input circuit is given by

$$C' = C_{gf} + (\alpha + 1)C_{gp}. \quad . \quad . \quad . \quad . \quad . \quad . \quad (61)$$

Hence any factor which affects the voltage amplification of the triode circuit, at the same time affects the capacity of the input circuit of the triode.

If then we measure the input capacity of the triode, as a function of the resistance in the plate circuit, we should find this capacity to increase about as indicated in Fig. 126; the input capacity approaches the value $(C_{gf} + (\mu + 1)C_{gp})$ as an asymptote. As the resistance R is increased, the loss in C_{gp} condenser circuit is increased, hence the conductance of the input circuit may be expected to increase as R is increased. The loss in the resistance, due to the charging current flowing through it, is comparatively small, compared to the loss inside the tube itself, hence we should not expect a very rapid increase in conductance as R is increased, and such is found experimentally to be the fact.

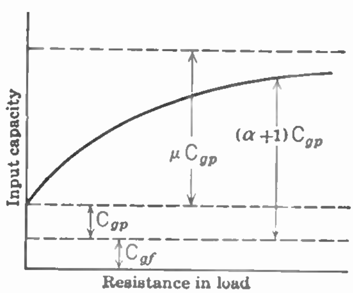


FIG. 126.

If an inductive reactance of low power factor is used in the plate circuit, in place of the resistance R, an even more rapid rise in the value of the input capacity takes place. This follows from the fact that α is greater if a given number of ohms reactance is used in the plate circuit than if the same number of ohms of resistance is used. For resistance we have

$$\alpha_r = \mu \frac{R}{R + R_p} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (62)$$

and for a reactance in the plate circuit we have

$$\alpha_x = \mu \frac{X}{\sqrt{R_p^2 + X^2}} \quad . \quad . \quad . \quad . \quad . \quad . \quad (63)$$

When $R = X$ it is evident that α_x is greater than α_r.

With the inductively reactive plate circuit another interesting fact is discovered. As the value of X is increased the conductance of the input circuit *decreases,* instead of increasing; as it did for the increasing R. Furthermore, the input conductance *may actually become negative* for a certain range in values of the reactance X, showing that instead of increasing the damping of the circuit to which it is attached, the input circuit of the triode actually *decreases* its damping, making the circuit more selective than it is without the triode connected. The conductance curve which may be obtained from this part of the test looks about as shown in Fig. 127. The amount of negative conductance which can be obtained depends upon the power factor of the reactance, and the amount of capacity between the grid and plate of the triode. For a

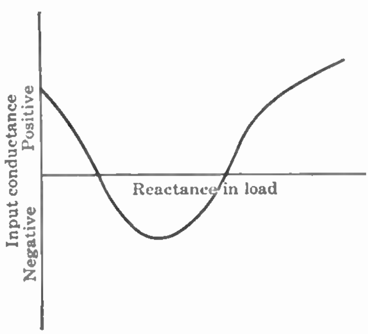


FIG. 127.

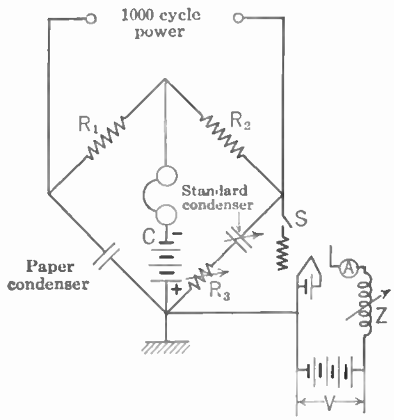


FIG. 128.

given circuit arrangement it increases with the frequency at which the measurement is made.

To measure the input capacity and conductance of a triode several methods are available. The method to be used in this test employs a capacity bridge of the differential type, as was used in Exp. 11, for measuring the specific inductive capacity and phase angles of dielectrics. The standard condenser is to be balanced against a paraffin-paper condenser (or similar kind) of about 1000 $\mu\mu f$ capacity. Of course, a resistance will be required in series with the standard condenser to balance the bridge because the power factor of the standard condenser is much lower than that of the paper condenser. The triode is wired as shown in Fig. 128, the small switch S being placed close to the grid. The battery C serves to make the grid as negative as desired. The impedance in the plate circuit should be variable, or adjustable in about ten steps, to give values from about one-quarter the resistance of the plate circuit of the triode to a value about three times as large as the plate resistance.

After all the conditions of the triode circuit have been properly adjusted with switch S closed, this is opened and the bridge is carefully balanced. (If the free grid results in a dangerous plate current, another switch can be put in the plate circuit, and this can be opened before switch S is opened.) Then switch S is closed and the standard condenser is decreased to again obtain an accurate balance. Of course, it will generally be necessary to readjust R_3 to get a good balance.

The capacity of the input circuit is obtained directly from the required change in setting of the standard condenser, and the resistance (or conductance) of the input circuit can be calculated from the change in condenser setting and change in R_3, as was analyzed in Exp. 11.

Carry out measurements using both resistance and inductive reactance in the plate circuit, obtaining about ten points for each curve. Using the special bridge of Exp. 12 measure the various capacities of the triode used.

From the known value of R_p (obtained from the readings of the c.c. meters in the plate circuit and the relations between R_{op} and R_p obtained in Exp. 24) and known value of impedance used in the external plate circuit, calculate the capacity of the input circuit from eq. (61) to check with the measured values.

EXPERIMENT 26

Object.—Study of the triode as a power converter, using separate excitation on the grid circuit. The effect of various adjustments on the amount of a.c. power obtained and the efficiency of conversion.

Analysis.—As has been noted in some of the previous experiments, if an alternating voltage E_g is impressed between the grid and cathode the effect in the plate circuit is the same as though a voltage of μE_g had been introduced in this circuit. The amount of alternating current set up in the plate circuit is given by the expression

$$I = \mu E_g/(R + R_p) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (64)$$

where R is the external resistance and R_p is the internal resistance (a.c.) of the plate circuit. This current will develop power equal to I^2R in the external circuit and it can easily be shown that this power is a maximum when $R = R_p$. This, it will be recalled, is a general law for electric generators; they deliver maximum power to their load when the internal and external resistances are equal. In testing this relationship with an ordinary generator the generated voltage must be lowered to a small fraction of its rated value or the machine will be burned out.

In trying out tests of this kind on triode converters the student is confronted with a new condition—the more power the triode delivers to its load the less is the heat generated in the triode itself. This is just the contrary effect to that which exists when working with ordinary generators.

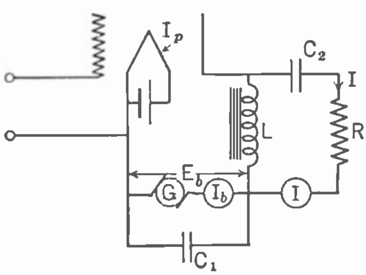


Fig. 129.

The B battery, or the c.c. generator which takes its place, delivers a certain amount of power to the triode and its associated circuit; part of this power appears as heat on the plate of the triode and the rest is delivered as alternating current power to the load circuit. In Fig. 129 the machine G delivers an amount of power E_bI_b. If there is no alternating voltage on the grid circuit (that is, no excitation) the current in the plate circuit will be uniform, and with the exception of the small amount of RI_b^2 in the choke coil L, all the power of machine G will appear as heat on the plate of the triode.

If now an alternating voltage E_g is impressed on the input circuit, the plate current must fluctuate; it is best to keep the fluctuations out of the generator G, so this is shunted by condenser C_1. No fluctuations can flow through choke coil L, so only the steady component of I_p (that is, I_b) flows here, and the rest of the plate current flows through the circuit $C_2 - R$, as the alternating current, I.

If the value of I_b remains constant when E_g is impressed (it is possible so to adjust conditions of plate voltage and grid bias that this is true) the power delivered by G is the same. But with E_g impressed, an alternating current power equal to I^2R is generated in the load circuit of the triode, hence the heat generated at the plates must diminish by this amount. The plates actually cool down when the triode starts to deliver power to its load circuit.

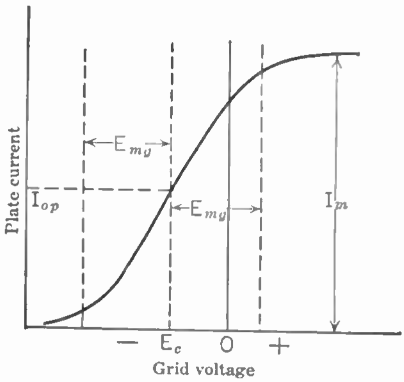


FIG. 130.

Let us suppose the characteristic $I_p - E_g$ curve of the triode is given by Fig. 130; the lower part of the curve and upper part are symmetrically located with respect to the grid bias E_c, and this is the proper bias to use. Here the plate current I_{op} is half of its maximum possible value, the total emission I_m. For comparatively small fluctuations in grid voltage the relation between E_g and I_p is linear, so that the fluctuation in plate current (that is, the alternating current of the plate circuit) is of the same form as the grid voltage E_g.

If, however, the grid excitation exceeds the value E_{mg} (Fig. 130), the plate current fluctuation becomes distorted and the a.c. output of the triode has harmonics, generally undesired. This is ordinarily the condition which limits the possible output of any triode.

With a resistive load circuit the plate voltage is a minimum at that time in the cycle when the plate current is a maximum, and this occurs when the grid has its maximum positive value (or minimum negative value). This consideration leads to a simple proposition regarding the proper bias for the grid, and proper grid excitation.

The grid bias should be sufficient to bring the plate current to the middle point of its characteristic curve. This current, multiplied by the plate voltage, should not give a power greatly in excess of the safe power rating of the plate.

The grid excitation should be sufficient to make the grid swing somewhat positive; just how much depends upon the type of triode. Referring to Fig. 130, it is seen that the relation between grid voltage and plate current is linear in the operating range; the same relation holds between plate voltage and plate current. To give a plate current twice the value of I_{op} would then require a voltage $2E_b$. But at the time when this high value of plate current is desired the plate voltage is so low as to be negligible, so the grid voltage must be sufficient to cause this plate current to flow, and as the grid is μ times as effective as the plate, the required peak value of grid excitation (E_{mg} in Fig. 130) must be equal to $\frac{2E_b}{\mu}$. Hence the value of required grid excitation in effective values is $\sqrt{2}E_b/\mu$.

This value of grid excitation will cause the plate current to fluctuate over the curved portions of Fig. 130, so there will be appreciable harmonics in the alternating current output.

In setting up the circuit of Fig. 129 certain precautions are to be observed. The filament current should be kept well within its rating, and the power used on the plate must be kept within proper limits. The reactance of the choke coil L should be at least ten times as great as the resistance of the plate circuit R_p. The resistance R should be adjustable in about ten steps from a maximum of about three times R_p to about one-quarter R_p. The reactance of C_2 should be small compared to the smallest value of R to be used in the test.

It will generally be necessary to use an iron core coil for the choke L. In this case care must be exercised that the normal plate current I_{op} is not sufficient to saturate the core. Generally the high-voltage winding of a small lighting transformer will be suitable. Its reactance (at its rated frequency) can be approximated by assuming the exciting current to be 10 per cent of its rated current and considering the no-load impedance to be all reactance. Thus a 1.5 KVA 2200 volt winding would have an exciting current of about 0.07 ampere, so its reactance is about 30,000 ohms.

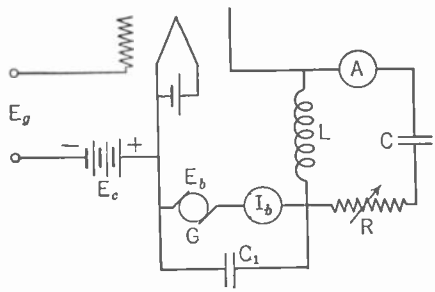


FIG. 131.

This core would saturate with a continuous current about twice the maximum value of exciting current, or about 0.2 ampere.

In general a triode feeds its power into a tuned load circuit, so this arrangement is also to be investigated in this test. The arrangement is shown in Fig. 131, where the iron core choke has been replaced by the air core inductance L, and the large capacity condenser C_2 has been replaced by one sufficient to give resonance (with L) for the frequency of the exciting voltage, E_g. A variable non-inductive resistance R serves to adjust the resistance of the load circuit.

As the proper selection of values for this circuit is somewhat more difficult than for the foregoing one, typical values will be approximately calculated. We will suppose a 250-watt tube, rated to take 2000 volts on the plate, and normal plate current 0.12 ampere with zero grid bias. The plate resistance $R_p = 2000/(2 \times .12) = 8000$ ohms. The excitation voltage on the grid is to be 2500 cycles and the μ of the tube is 40.

The proper maximum grid excitation voltage (maximum value) is $4000/40 = 100$, or an effective grid excitation of 70 volts. The grid will of course draw some current with this amount of excitation.

We will suppose that a low-resistance coil of $L = 0.05$ henry is to be used. The reactance of this coil at 2500 cycles is 750 ohms. A reasonable power factor for this coil would be 0.03, so its resistance, at 2500 cycles, would be about 20 ohms.

A condenser of about $0.1\mu f$ will give resonance, and as the voltage across it will have a peak value of nearly 2000 volts, the condenser should be of mica, or similar strong dielectric. At resonance a coil of 0.05 henry and 20 ohms resistance in parallel with a condenser of $0.1\mu f$ and negligible loss will show a resistance of about 24,000 ohms.

The variable resistance R should have a maximum value of about 200 ohms so that the resistance of the parallel circuit can be reduced below that of the triode, namely, 8000 ohms. With 180 ohms added in the circuit, making a total resistance of 200 ohms, the resistance of the parallel circuit is 2500 ohms. Thus by varying the rheostat the resonant load circuit can be made to have any resistance between 24,000 and 2500 ohms, both higher and lower than the plate circuit resistance of the triode.

The ammeter for measuring the alternating current in the tuned circuit can well be of the hot wire type, and its proper range calculated by assuming that about 100 watts of power will be delivered to the tuned circuit when its resistance is equal to that of the triode. This will require a resistance in the tuned circuit of about 100 ohms, hence the current in the tuned circuit will be about 1 ampere.

An error will occur in the reading of the ammeter by which the alternating current output of the tube is read, no matter in which arm of the parallel circuit it is placed. This ammeter should read the effective value of the fundamental component of the current, that is, the effective value of the current which has the same frequency as the voltage impressed on the grid.

Now the tube always gives some harmonic currents, as well as the fundamental, and the magnitude of these higher frequency currents increases rapidly with the output. These harmonic currents flow through the condenser arm of the load circuit, and so act to increase the reading of the ammeter if it is placed in this branch, as shown in Fig. 131. As the amount of these upper harmonic currents is not known it is impossible properly to correct the ammeter reading so as to get the magnitude of the fundamental frequency current.

If the output ammeter is placed in the inductance branch of the parallel circuit the accuracy of its reading will not be affected by the upper harmonic output of the triode, because the high reactance of the coil will practically exclude them. However, in this branch of the parallel circuit, the steady component of the plate current flows, in addition to the alternating current of fundamental frequency, so the ammeter here will read too high by an amount depending upon the ratio of this continuous current to the alternating current.

This error can, however, be calculated, because the continuous current ammeter I_b (Fig. 131) does measure the average plate circuit current. Hence the desired value of fundamental alternating current can be obtained from the reading of the two meters. Suppose the hot wire ammeter, when placed in series with coil L of Fig. 131, reads 0.65 ampere and the ammeter I_b shows the average plate current to be 0.2 ampere. The fundamental alternating current is then $\sqrt{0.65^2 - 0.2^2} =$ 0.616 ampere.

Although the circuit assumed above was analyzed for a frequency of 2500 cycles it is of course possible to run the test at higher or lower frequencies, as the laboratory apparatus permits. At the lower frequencies difficulty may be encountered in matching the resistance of the load circuit to the tube resistance, however. Thus suppose the test is to be run at 100 cycles. A good coil of 0.6 henry is available, having a resistance of only 12 ohms. (This is about as good a coil as can be built with 75 lb. of copper cable.) About 5 microfarads of capacity will be required to give resonance, and the circuit will show a resistance (at resonant frequency) of only about 10,000 ohms, only slightly greater than the triode resistance of 8000 ohms. Thus if the test is to be run at the lower frequencies a triode of lower plate circuit resistance is advisable, so that the load may offer a resistance two or three times that of the tube itself. Of course the lower values of load circuit resistance are easily obtained by putting a non-inductive rheostat in the tuned circuit; the more resistance the rheostat has, the less is the resistance of the parallel circuit, as shown in Exp. 7.

If the test is to be run at 100 cycles, with the coil and condenser mentioned above, the plate circuit of the triode should have a resistance of about 2500 ohms. This will generally mean a tube of lower μ than the one assumed before. But a lower μ means that a higher exciting voltage must be used, so that the 100-cycle source must probably be connected to the grid through a step-up transformer.

As the source of excitation, whether generator or vacuum tube oscillator, is sure to have some upper harmonics in its wave form, it is well to arrange the exciting circuit to eliminate them as much as possible. A suitable arrangement of exciting circuit, if generator is used,

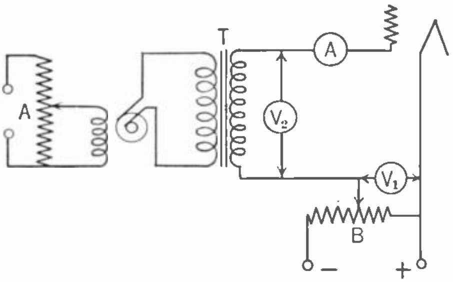


Fig. 132.

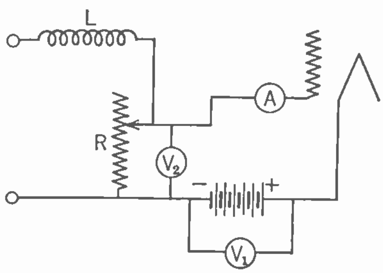


Fig. 133.

is shown in Fig. 132. The input circuit of the triode is excited from a sine wave machine, so no precautions are necessary to eliminate upper harmonics. The field of the alternator is excited from potentiometer A, and a second potentiometer B serves to get the requisite bias. Transformer T raises the alternator voltage to the several hundred which a power tube with low μ will probably require. Voltmeter V_1 reads the grid bias, and voltmeter V_2 reads the exciting voltage.

In case a vacuum tube oscillator is used for excitation, the arrangement of Fig. 133 is applicable. The resistance R has a maximum value somewhat greater then the value which gives the maximum output of the oscillator. The choke coil L has a reactance about equal to the maximum value of R. The voltmeter V_2 for reading the excitation voltage will have to be a sensitive thermocouple meter, of high resistance, otherwise it may take more power than the oscillator can give. If the resistance R is a decade resistance box (which it may well be) a thermocouple ammeter may be used to read the current through it, and so the exciting voltage calculated from this current and the known value of R.

The value of R (and consequently the output resistance of the oscillator) should be as low as possible, because the grid draws some current under certain conditions of the test, and this grid current, flowing through resistance R, may affect the wave form of the voltage impressed on the grid. The c.c. ammeter A, in the grid circuit, will read a few milliamperes when the conditions are adjusted for maximum output of the triode, perhaps 5 mils for a 5-watt tube, and perhaps 25 mils for a 250-watt tube. This reading is the *average* grid current; the maximum may be ten times the average. This maximum grid current, flowing through the resistance R, must give an RI drop small compared to the value of the exciting voltage, or the exciting voltage will be flat-topped on the positive alternation, i.e., during that part of the cycle when the grid is drawing current.

The ammeter A will read zero for all adjustments which give a reading of V_2 less than $1/\sqrt{2}$ the reading of V_1 (Figs. 132 and 133), because it is only when the grid swings positive that it draws an appreciable current.

With a resistive load (Fig. 131) make sufficient tests to find out how the output of the triode varies with filament current, grid excitation, grid bias, plate voltage, and load resistance. Make one run in which grid excitation and grid bias are simultaneously varied to keep V_2 just 0.7 of V_1 (of Figs. 132 and 133), filament current, plate, voltage, and load resistance being held fixed at normal values.

Make one run with the tuned load circuit, keeping filament current, plate voltage, grid bias, and grid excitation at normal values, no resistance added in tuned circuit, varying frequency both above and below resonant value. Holding frequency at resonant value, and other factors at normal, vary the resistance in the tuned circuit in about ten steps, to give load circuit resistances both greater and less than the triode plate resistance.

Measure the resistance of coil, condenser, and hot wire ammeter in series by connecting them to the source of excitation and measuring volts and amperes when resonant frequency voltage is impressed.

This resistance is to be added to the known values of resistance in the rheostat, to give the actual tuned circuit resistance.

Plot proper curves to show the various relationships investigated in these tests.

Object.—Study of a triode arranged for producing alternating current by self-excitation. Effect of various circuit adjustments on possibility of setting up oscillations and on amount of power the triode delivers.

Analysis.—The results obtained in the foregoing experiment show that if the grid-filament circuit of a triode is properly excited, and if the external plate circuit is resistive and of about the same resistance as the internal plate circuit resistance, an amount of alternating current power can be delivered to the load circuit equal in amount to possibly 50 per cent of the continuous current power supplied to the plate circuit by the *B* battery, or its substitute. The figure of 50 per cent is rather optimistic, from 25 to 40 per cent being more usual.

To get this amount of power the grid must be sufficiently excited, and the fact that the load circuit must be resistive really specifies that the fluctuations in plate potential (i.e., the alternating voltage on the plate) must be about 180° out of phase with the voltage on the grid.

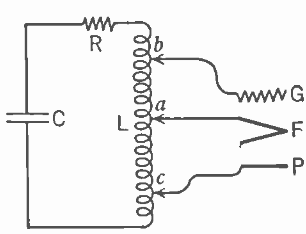


Fig. 134.

It does not require very much power to excite the grid, so the idea of exciting the grid from the load circuit of the triode, instead of using a separate source of excitation, is more or less self-suggestive. And experiment shows the applicability of the idea; summarized briefly it may be stated as follows.

If a low-resistance circuit made up of coil and condenser (or coils and condensers) is connected to a triode in such away that when alternating current flows in the circuit the plate and grid fluctuate in voltage with respect to the filament, sufficiently, and opposite in phase, the apparatus will set up self-sustained oscillations in the L–C circuit, as long as the filament and plate circuits are supplied with the requisite power.

This idea is the fundamental basis of all self-excited oscillatory circuits, so it will be analyzed more in detail. In Fig. 134 there is shown a circuit made up of coil *L*, having resistance *R*, in series with the condenser C, of negligible series resistance. This circuit, if it oscillates at all, will oscillate at the frequency fixed by the relation $f = \frac{1}{2\pi\sqrt{LC}}$. The power used in the circuit will be I^2R, where I is the effective magnitude of the alternating current.

All batteries have been omitted for simplicity; the filament has been attached to the mid-point of the coil, and the grid and plate have been attached to sliding contacts, one on either side of the filament. From this connection scheme it is evident that if alternating current flows in the coil the points b and c will fluctuate in potential with respect to point a by an amount depending upon the amount of current in the circuit, and upon the amount of inductance included between $a - b$ and $a - c$, respectively. Furthermore, the point b will go up with respect to point a, when point c is going down with respect to a. This means that the variations in potential of the grid and plate, with respect to the filament, will be opposite in phase.

We will now assume actual constants for the circuit, to calculate whether or not the triode can sustain oscillations. We will assume that the rating of the tube is 5 watts output, and that input to the plate circuit, from the B battery, is 15 watts. A 5-watt output therefore assumes a $33\frac{1}{3}$ per cent efficiency, a reasonable figure.

The coil has 300 microhenrys inductance and the condenser has 500 micro-microfarads capacity. The natural frequency of the circuit is then 410 kilocycles. The reactance of the coil is 775 ohms. A reasonable power factor for the coil is 2 per cent, so we assume its resistance is 14 ohms. (This is of course the high-frequency resistance; the c.c. resistance would not be more than 1 or 2 ohms.)

We assume that the plate voltage is 300 volts and the μ of the tube is 5.

From previous analysis we know the plate voltage must fluctuate from nearly zero to nearly 600 volts, and that the fluctuations of the grid potential must be at least $2 \times \frac{300}{5} = 120$ volts. So the effective plate voltage (a.c.) must be $300/\sqrt{2} = 210$ volts, and the effective voltage on the grid must be $120/\sqrt{2} = 84$ volts.

Now if 5 watts are supplied to the oscillating circuit, and this has 15 ohms resistance, the current must be $\sqrt{5/15} = 0.58$ ampere.

The reactance from the mid-point on the coil L (Fig. 134) to either end is about 387 ohms. The reactance drop is then $387 \times .58 = 225$ volts. Therefore the contact c (Fig. 134) should be moved practically to the bottom of the coil, and the contact b, to give grid excitation, would be moved up on the coil somewhat more than one-third of the way toward the top. This position would give the required 84 volts for grid excitation.

If now the resistance of the oscillatory circuit is increased to 60 ohms, by adding 45 ohms in series with the coil, the current would decrease to $\sqrt{5/60}$ or 0.29 ampere. The drop over one-half of the coil would now be only 387 × 0.29 or 113 volts. Hence even if the plate tap is moved down to the bottom of the coil it could not make the plate voltage fluctuate more than $113\sqrt{2} = 158$ volts, whereas it must fluctuate about 300 volts if the triode is to deliver 5 watts to the oscillating circuit.

Hence the circuit, with 60 ohms resistance, could not oscillate with the conditions assumed. However, if the triode would operate safely with more plate volts, or less grid bias, so that it could deliver say 10 watts to the oscillating circuit, the outfit might oscillate even with the 60-ohm resistance. One way of helping the triode to oscillate would be to move the connection *a* up from the center, thus making more than half the coil available for furnishing plate voltage. The grid needs less excitation than the plate, so point *a* might well be moved to a point about one-third from the top of the coil.

Another circuit is shown in elementary form in Fig. 135. Here the tuning condenser is shown as the two condensers C_1 and C_2 in series. When oscillations occur the plate is excited by the voltage drop across C_2, and the grid is excited by the drop across C_1. As the reactance of these two condensers, and the frequency and magnitude of the oscillation, can easily be calculated, the possibility of this circuit oscillating can be checked exactly as was done for the split coil circuit of Fig. 134. The L-R-C_1-C_2 circuit of Fig. 135 could not possibly work without other circuit arrangements added (in addition to batteries, etc.), because the filament is isolated from the plate, so that there could be no electron flow from filament to plate, and without this of course the triode could not function. Furthermore, even if the plate circuit, for the continuous current, was completed by the battery and choke coil shown in Fig. 135, the arrangement would still be inoperative because the grid would be at the same positive potential as the plate; an extra grid condenser and grid leak to filament would be required to make the circuit operable.

The student will soon learn that apparently minor changes in circuit arrangement make the difference between success and failure in making triodes oscillate. One such point is illustrated in Fig. 136.

A generator and choke coil are used to furnish plate voltage; the filament circuit is grounded, generally intentionally, but even when not intentionally the filament batteries or filament power supply practically ground it, in so far as high-frequency oscillations are concerned.

Now with the plate generator G and choke L connected as shown, the generator would have to go up and down in potential, at the high frequency, as much as the plate of the triode. But the generator frame is generally grounded more or less, and the capacity between armature wires and armature core is sufficient to practically ground the armature

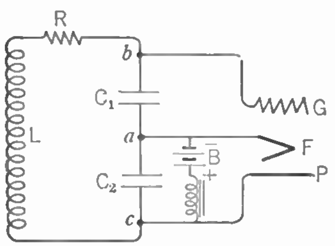


FIG. 135.

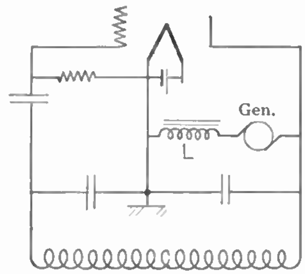


FIG. 136.

windings so far as radio frequency currents are concerned. Hence the plate of the triode is really connected to ground (for high frequency) through the insulation of the armature windings, and probably no oscillations will occur. If they do occur it is quite likely that the armature winding insulation will soon break down because of the excessive dielectric loss caused by the high-frequency currents.

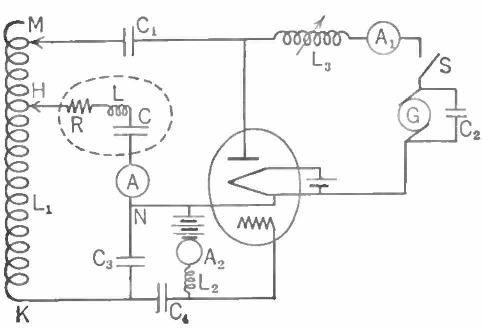


FIG. 137.

However, if the positions of the choke coil L and generator G are interchanged, placing the generator next to the grounded filament, the circuit will probably oscillate without any trouble. However, in all circuits of this kind it is well to put a by-pass condenser across the generator G, to keep even small high-frequency currents out of the armature windings. Even if a dry cell battery is used for plate supply, a by-pass condenser around the battery will generally increase the power output to a small extent.

In Fig. 137 is shown the complete wiring diagram for such an oscillating circuit as has been frequently used for portable radio telephone sets. For a 5-watt tube the proper values, for a frequency of about 1000 kc., are as follows:

The generator G is 25-watt capacity and 300 volts. It is shunted by a 1-microfarad capacity, C_2. Radio frequency choke coil L_3 is variable from zero to about 1000 microhenrys. Ammeter A_1 is of 100 milliampere capacity. Condenser C_1 is an insulating condenser of about $0.1\mu f$ to keep the 300-volts supply away from coil L_1. This coil L_1 has two variable or adjustable contacts M and H. Its total inductance is about 100 microhenrys. The resistance R, coil L, and condenser C, represent a typical small antenna, R being 10 ohms, L being $15\mu h$, and C about $200\mu\mu f$. Ammeter A, of the hot wire or thermocouple type, has a range of one ampere. Condenser C_3 is variable, or adjustable in steps, from $50\mu\mu f$ to $1000\mu\mu f$. Coil L_2 is a small (physically) radio frequency choke coil, of reactance large compared to the reactance of C_3 when set at its smallest value. C_4 is a grid condenser for which various sizes can be substituted. The ammeter A_2 is of about 10 milliampere range.

The oscillating circuit is made up of that part of coil L_1 between K and H, coil L, and condensers C and C_3 in series. The frequency of the alternating current can be quite closely calculated from the values of these quantities.

The plate circuit of the triode supplies the alternating current power to the oscillating circuit across points M and N, and for the frequency generated the power factor of the oscillating circuit between points M and N is unity, as we know it should be to get much power from the tube. The amount of resistance which the oscillatory circuit shows across points M–N may be varied by moving the contact M. If the actual resistance of the oscillatory circuit is, say, 15 ohms, the apparent resistance across points M–N may be as much as 10,000 ohms. By sliding contact M downward on the coil L_1 this apparent resistance may be diminished to practically zero. This circuit was analyzed in Exp. 7.

If the triode is to generate power the plate must be able to go up and down in potential with respect to the filament; this is the function of choke coil L_3. It will be found that even if all the other conditions are right, no oscillations will occur until a certain minimum value of L_3 is used. As this value is exceeded the amount of power supplied to the oscillatory circuit increases slowly, provided the resistance of L_3 is negligible compared to the plate circuit of the triode.

It will be found that this circuit will generally oscillate at once if C and C_3 are made equal and M and H are put close together. The grid bias should of course be the rated value for the triode.

Study the operation of this oscillatory circuit varying one factor at a time; the dummy antenna is to be kept fixed as this would not be changeable, of course, if the triode was furnishing power to an actual antenna.

It will be noticed that changing C_3 affects primarily the grid excitation, and secondarily the frequency of the alternating current; secondarily because it will be found that for best power output the capacity of condenser C_3 is considerably greater than that of condenser C so that C has more effect on frequency than C_3.

Shifting the contact H changes the frequency proportionately; in fact, this is the frequency adjustment for the set. Shifting contact M changes the apparent resistance of the oscillating circuit, as it appears to the plate circuit of the tube. For best output the position of contact M must be changed for every condition that affects the resistance or frequency of the tuned circuit, or the resistance, R_p, of the triode. For best conditions the resistance of the oscillating circuit, as it appears between the points M and N, should be equal to R_p.

Changing the value of C_4 will have negligible effect until it becomes very small; if it is changed from one microfarad to one millimicrofarad practically no effect is produced on the oscillatory circuit, but if it is made only a few micro-microfarads the excitation supplied to the grid becomes so small that the triode cannot oscillate.

If the choke coil L_2 is made too small the circuit cannot oscillate, as here also the grid will get insufficient excitation.

After the circuit has been investigated, substitute a grid leak for the grid choke coil L_2 and the biasing battery. In trying various values of grid leak, precautions must be observed, with an oxide-coated cathode tube especially, when high values of resistance are used. It may be that, with high values of leak resistance, the grid swings positive when oscillations start, and this will result in such a large increase in plate current as to ruin the tube. This condition can occur either when there is gas in the tube (the oxide-coated filament tube generally has more gas than the tube with tungsten filament), or when sufficient secondary emission from the grid occurs.

With this arrangement try the effect on wave length, power, plate current, etc., of varying the capacity of the grid condenser, and the resistance of the leak.

Caution.—When the grid leak is removed from its clip, to change for another, the grid is "free" and may swing positive and ruin the tube. Hence it is best to open the plate circuit when the leak is being changed.

In all the tests called for above, the various quantities, plate current, grid current, oscillating circuit current, wave length, etc., should be read for each adjustment. Qualitative conclusions can then be drawn regarding the best adjustment of the circuit.

EXPERIMENT 28

Object.—Measurement of resistance and reactance of a tuned circuit, as this is affected by the regenerative action of a triode. Study of the action as the mutual induction between input and output circuits is varied. Effect of varying the impressed frequency.

Analysis.—In such a circuit as that shown in Fig. 138 the resistance, and possibly the reactance, may be expected to vary as the coupling between L_1 and L_2 is changed, for reasons which can be deduced without any exact mathematical analysis. Suppose the circuit is opened at some such point as A, so that the resistance and reactance of the L_2–C circuit can be measured in a Wheatstone bridge. When the measuring

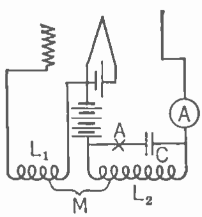


FIG. 138.

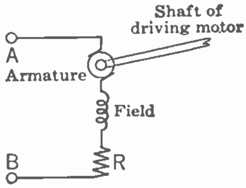


FIG. 139.

current flows through the L_2 coil, it will induce a voltage in L_1, thus affecting the grid potential. But we know that the current flowing in the plate circuit of a triode is controlled by the grid potential, so that a different current will flow through the L_2–C circuit than would flow if the coupling between L_1 and L_2 were not present. But if the current flow, for a given impressed voltage, is changed, we interpret this as a change in resistance or reactance (or both) of the circuit in which the current is flowing.

Of course, we know perfectly well that the true conductor resistance of the circuit does not in any way change as the coupling of L_1 and L_2 is altered; for a given current flowing through it, there is just as much heat developed in coil L_2 whatever the coupling between L_1 and L_2 may be. However, when impedance is determined by bridge measurement, or by the ratio of voltage to current, the measured value will be found greatly dependent upon the grid-plate circuit coupling, and these measured values of R and L, hypothetical as they may seem, have real significance so far as current flow in the circuit is concerned.

A somewhat analogous case may be set up in a continuous current circuit, as shown in Fig. 139. A series generator is connected in series with the resistance R which is to be measured. The external characteristic curve of the series generator is nearly a straight line, as shown in Fig. 140. With 5 amperes flowing through it, the voltage across its terminals is 50 volts, and with 10 amperes flowing it is 100 volts. After this its voltage increases somewhat more slowly, owing to saturation of its magnetic circuit.

Now suppose that the actual resistance of R is 20 ohms. When 50 volts is impressed across the terminals A–B of Fig. 139, how much current will flow? Evidently the series generator will help to force current through the circuit so that more than $2\frac{1}{2}$ amperes will flow. By a cut-and-try method we can soon find the answer to be 5 amperes. Such a current will require 100 volts to force it through the 20-ohm resistance, but of this 100 volts 50 will come from the generator to add to the 50 impressed on terminals A–B. The resistance of the circuit would then be found to be 50 volts/5 amperes = 10 ohms. To check this answer we will increase the voltage impressed across terminals A–B to 100 volts. The current that will now flow is found to be 10 amperes. This current requires 200 volts to force it through the 20-ohm resistance; 100 volts is supplied by the voltage across A–B, and the other 100 volts is furnished by the series generator. The resistance is now found to be 100 volts/10 amperes = 10 ohms, the same as before. And so it would be found for any other value of impressed voltage if the current which flowed in the circuit was less than 10 amperes.

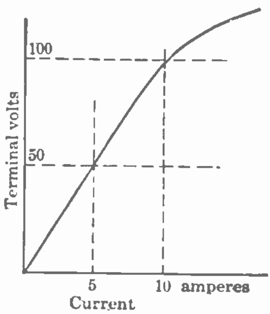


FIG. 140.

Now, as we know that the resistance of R is 20 ohms, but when it is measured in series with the generator it always measures 10 ohms, we reach the unavoidable conclusion that the generator is equivalent (so far as current flow is concerned) *to minus 10 ohms.* That is the generator may be considered merely as 10 ohms of *negative resistance.*

Continuing this analysis, suppose the actual resistance of R is decreased to 11 ohms, and 5 volts are impressed on terminals A–B. The current that flows now will be 5 amperes. Of the 55 volts required to force the 5 amperes through 11-ohm resistance, 50 is supplied by the series generator and 5 by the outside source. The resistance of the circuit is now found to be, using Ohm's law, 5 volts/5 amperes = 1 ohm.

If now the value of R is diminished to 9 ohms it will be found that about 12 amperes of current flow, even when the impressed voltage is zero! If the saturation of the iron of the magnetic circuit of the generator did not act to limit the voltage it can generate, the current would rise to tremendous values, being limited only by the increased resistance of the wires caused by temperature rise.

Thus, a circuit which has a negative resistance is an unstable one; current starts to flow as soon as the circuit is closed, and this current increases in magnitude until the circuit is destroyed, unless some saturation or other limiting effect acts as a stabilizer.

At this point in the analysis it is to be noticed that a circuit can show negative resistance only if it contains a source of electrical power, which generates a voltage *in phase with the current flowing in the circuit.* And if the negative resistance is to be constant in magnitude, the voltage generated by the source must be proportional to the current flowing in the circuit.

Now the circuit of Fig. 138 does offer the possibility of negative resistance, because current flow in the tuned plate circuit may actually generate in the circuit a voltage in phase with itself, through the combined action of the mutual induction between L_1 and L_2, and the grid control of plate current.

A careful analysis of this circuit by means of differential equation relationships,* shows that the resistance of the L_2–C circuit, as measured by a bridge, should be given by

$$R' = R_L + r + \frac{1}{\omega C^2 R_p} + \frac{\mu M}{C R_p} \quad . \quad . \quad . \quad . \quad . \quad (65)$$

in which

R_L = resistance of L_2
r = equivalent series resistance of C
ω = 2π times the impressed frequency
R_p = a.c. resistance of plate circuit of triode
μ = amplification factor of triode
M = mutual induction between L_1 and L_2

It will be noticed, from Fig. 138, that the resistance R_p is shunted directly across C, in so far as alternating current considerations are concerned, so that the term $\frac{1}{\omega C^2 R_p}$ is really the resistance in series with C which is equivalent to the resistance R_p in shunt with C. Hence, the

* See "Principles of Radio Communication." 2nd ed., p. 571 et seq.

first three terms of eq. (65) are all familiar to us; they represent the total positive resistance in the L_2–C circuit.

Now the value of M may be either positive or negative, in triode circuit analysis, according to the relative polarity of connections of L_1 and L_2. That is, if with the coils in a certain position we regard M as positive, then by merely reversing the connections of either one, the value of M must be considered as negative because the same increment of plate current will give voltage of opposite polarity on the grid; this will have an opposite effect on the plate current itself. In the differential equation analysis from which eq. (65) was derived, it was assumed that $e_g = -M\frac{di_p}{dt}$, which says that if a positive increment in plate current gives a positive grid voltage M is to be considered as negative, and, of course, vice versa.

If then the coils L_1 and L_2 of Fig. 138 are so connected as to give a negative M then it is possible for the value of R' to actually become negative. In either case, R' should vary linearly with M, either increasing or decreasing directly as M is altered. The two possibilities are shown in Fig. 141; with positive M, R' increased from the value $R_L + r + \frac{1}{\omega C^2 R_p}$ and with negative M it decreases, and actually becomes negative when M has absolute values great enough to make $\frac{\mu M}{CR_p}$ greater than $R_L + r + \frac{1}{\omega C^2 R_p}$.

With the ordinary condenser r decreases with frequency increase, and of course $\frac{1}{\omega C^2 R_p}$ decreases with frequency increase, so that if another series of measurements of R' are carried out at a higher frequency, curves similar to the dashed lines of Fig. 141 might be obtained. In making this test it is convenient to use such a frequency that the reactance of the L_2–C circuit is zero; a bridge with resistance arms only is then required. As the change in position of two coils requisite to obtain the different values of M may possibly change slightly the reactance of coil L_2, it is advisable to put a continuously variable condenser in parallel with C and use this condenser to keep the reactance of the L_2–C circuit equal to zero, as the test progresses. It is advisable when beginning this test to set this variable condenser at its mid-point and adjust the impressed frequency to give resonance with $M = 0$. Then as the test progresses the capacity may be either increased or decreased as required to give resonance, the impressed frequency being held constant.

A convenient set-up of apparatus is shown in Fig. 142. A frequency of about 1 kilocycle, a condenser of 1 microfarad, an inductance L_2 of about 25 millihenrys, are convenient values to use in this test. As the resistances will be low, a slide wire makes convenient ratio arms and permits rapid settings. The value of R will be about 10 ohms for a unity ratio bridge when there is no coupling between L_1 and L_2. As the coupling is increased, the value of the term $\frac{\mu M}{CR_p}$ may go up to 20 ohms or more, using coils permitting tight coupling, and ordinary triodes having a μ of 5 and R_p about 5000.

If it is desired to use a frequency other than that which gives reso-

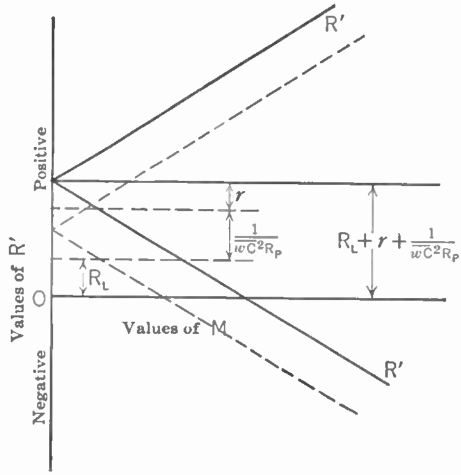


Fig. 141.

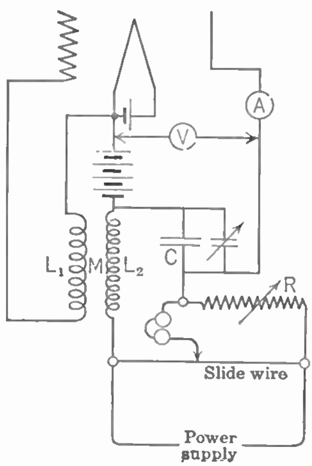


Fig. 142.

nance, the third arm of the bridge must contain a variable condenser or variable inductance, according as the frequency is lower or higher than the resonance value.

Make measurements of R' of the L_2–C circuit maintaining the resonant condition by varying the condenser in parallel with C, keeping the impressed frequency constant. Get measurements for about ten values of M.

Measure resistance and reactance of the L_2–C circuit after C has been increased to 10 microfarads, keeping impressed frequency same as before.

With $C = 1$ microfarad and a value of M almost large enough to give zero resistance at the resonant frequency, measure resistance and reactance of the L_2–C circuit for frequencies both above and below resonance. Get about ten measurements in a frequency range from 5 per cent above to 5 per cent below resonance.

Another circuit which is much used in radio apparatus is shown in Fig. 143; in this case the tuned circuit is connected to the grid instead of the plate, as was the case for the circuit previously analyzed. For this case the mathematical analysis shows the resistance of the tuned circuit to be

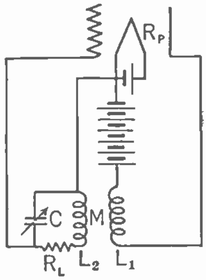


FIG. 143.

$$R' = R_L + r + \frac{R_p}{R_p^2 + \overline{\omega L_1}^2}\left(\overline{\omega M}^2 + \frac{\mu M}{C}\right) \quad . \quad . \quad . \quad (66)$$

in which the symbols have the same meaning as before. Although this expression looks entirely different from that of eq. (65) for the previous case, it can be shown to have almost the same form.

In the average circuit R_p is much greater than ωL_1, so that for

$R_p/(R_p^2 + \overline{\omega L_1}^2)$ we may write, without much error, $1/R_p$.

Then eq. (66) becomes

$$R' = R_L + r + \frac{\overline{\omega M}^2}{R_p^2 + \overline{\omega L_1}^2} R_p + \frac{\mu M}{C R_p}. \quad . \quad . \quad . \quad (67)$$

The third term is now seen to be the resistance which the plate circuit of the triode (the L_1 circuit) introduces into the tuned grid circuit, as was demonstrated in Exp. 4. For the conditions generally existing in triode circuits it may be written $\frac{\overline{\omega M}^2}{R_p}$, as stated above, so that we have

$$R' = R_L + r + \frac{\overline{\omega M}^2}{R_p} + \frac{\mu M}{C R_p}. \quad . \quad . \quad . \quad . \quad . \quad (68)$$

The measured values of R' will now be of somewhat different form from those obtained for the previous circuit, because of the form of the third term. In the previous case it was independent of M, whereas in this one it increases with the second power of M. Hence the two possible forms of R' will be as shown in Fig. 144. It will be noticed that if M is increased too much, the value of R' after having passed through a series of negative values again becomes positive because of the increasing effect of the term $\frac{\overline{\omega M}^2}{R_p}$.

R' will evidently return to the positive region slightly before $\frac{\overline{\omega M}^2}{R_p}$ becomes equal to $\frac{\mu M}{CR_p}$. This occurs when the absolute value of M becomes equal to $\frac{\mu}{\omega^2 C}$. Supposing that we use the same values of μ, R_p, ω, and C as before, we find that this value of M is practically impossible to obtain, so that actually the values of R' obtained experimentally have about the same form as those obtained in the former tests. If, however, the capacity is increased to about 10 microfarads (this will require a variable inductance in the third arm of the bridge to obtain a balance), then the form of R' obtained experimentally will have about the form shown in Fig. 144. The term $\frac{\overline{\omega M}^2}{R_p}$ will be equal to the term $\frac{\mu M}{CR_p}$ for a value of M which is reached with $L_1 = L_2 = 0.025$ henry coupled about 50 per cent.

Make a series of measurements of the effective resistance of the L_2–C circuit of Fig. 143, keeping the circuit tuned, as was done in the first part of this test, by a small variable condenser in parallel with C, the impressed frequency being held constant. Make measurements for about ten values of M. Then with the same impressed frequency measure the resistance and reactance of the L_2–C circuit when C has been increased to 10 microfarads.

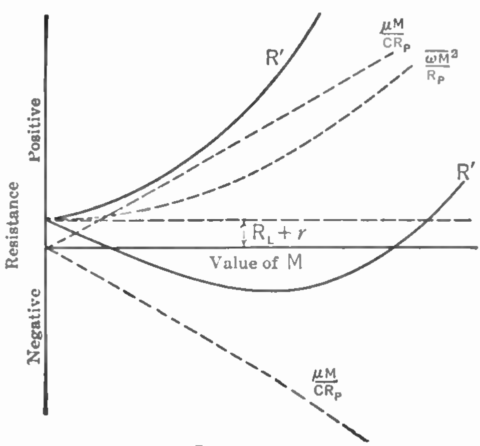


Fig. 144.

If time permits, with a value of M almost large enough to give zero resistance at the resonant frequency (with $C = 1$ microfarad), measure values of resistance and reactance of the L_2–C circuit for frequencies both above and below the resonant frequency in the same range as for the first part of the test.

EXPERIMENT 29

Object.—Measurement of coil resistance at radio frequencies. Various sizes of wire. Effect of using a tapped portion of a large coil.

Analysis.—To make an accurate determination of the resistance of a coil or condenser at radio frequency is a task requiring considerable skill and experience; results can be obtained which are repeatedly duplicated (with the same set-up), thus giving the student confidence in his results, yet when the apparatus is slightly rearranged entirely different results will be obtained.

Resistance measurements cannot be made by wattmeter and ammeter, as is generally done at power frequencies; no wattmeters for very small power, at radio frequencies, have been developed. Specially constructed bridges have been used for frequencies of several hundred kilocycles, but apparently the bridge is not reliable for frequencies measured in megacycles.

The method which is practically always used is based on the simple idea that in a resonant circuit the current is equal to the voltage divided by the circuit resistance, which is, of course, unknown. Instead of measuring the voltage impressed on the circuit (generally difficult to do as the circuit is arranged) this is held constant while a known non-inductive resistance is put in series with the circuit being measured. Enough of this known resistance is inserted to reduce the current to one-half; evidently then the amount of resistance added is equal to the resistance of the circuit itself.

Simple as this scheme seems, considerable difficulty is encountered in carrying it out. The circuit must be maintained in resonance within a small fraction of 1 per cent, the voltage impressed on the circuit must be exactly the same when the ammeter readings are taken, the same current must be flowing throughout the whole circuit, and the resistance of the circuit itself must be independent of the amplitude of current flowing in it. These points will now be taken up separately.

Let us suppose that a coil of 200 microhenrys is being measured, at one megacycle. Its resistance will be about 10 ohms at this radio frequency, whereas it would be only about 1 or 2 ohms for continuous current. At 3 megacycles its resistance would be from 25 to 50 ohms, as coils are ordinarily built. The reactance of the coil is about 1200 ohms at one megacycle, and of course the condenser in series with the coil to give resonance must also have 1200 ohms reactance.

Now suppose the frequency impressed on the circuit accidentally increases only $\frac{1}{2}$ of 1 per cent; what change will take place in the magnitude of the current? The reactance of the coil will increase by 6 ohms and the reactance of the condenser will decrease by 6 ohms, thus giving a circuit reactance of 12 ohms. The impedance of the circuit (which should be resistive only) now consists of 12 ohms reactance and 10 ohms resistance, and is found to be over 15 ohms. Thus the current in the circuit is decreased to 65 per cent of its proper value, by a frequency change of only $\frac{1}{2}$ of 1 per cent, and it is solely upon the current readings that the resistance determinations are to be made.

Even if the frequency change is only one-tenth of 1 per cent, the reactance introduced is 2.4 ohms, thus changing the current reading in the circuit by 3 per cent. With the ordinary wave meter it is impossible to even read the frequency with this precision.

To avoid the very large error which the frequency variation may introduce, the test circuit must be carefully retuned each time before a current reading is taken.

A very slight rearrangement of the wires used for connecting the parts of the circuit in series will change the inductance of the circuit by 1 microhenry, and 1 microhenry, at 1 megacycle, has a reactance of 6 ohms, more than 50 per cent of the resistance being measured!

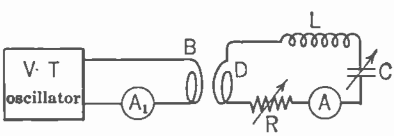


Fig. 145.

The constancy of the voltage induced in the test circuit must be determined indirectly; it cannot be directly measured. The arrangement of apparatus is conventionally shown in Fig. 145. The power is always obtained from a triode oscillator, generally of the self-excited type. A small coil B in series with the oscillating circuit of the power triode serves to induce radio frequency voltage in small coil D which is in series with the circuit being measured. The test coil L is in series with the variable condenser C, thermocouple ammeter A, and variable resistance R.

Now it was shown in Exp. 5 that in such an arrangement as this a very high resistance may be introduced into coil B by the presence of the circuit L–C–R–D. Furthermore, the amount of the resistance introduced in coil B varies inversely as the resistance of the test circuit. But it was found in Exp. 27 that if the resistance of the oscillating circuit of a self-excited triode was varied, the current in the circuit also varied. Hence as the resistance R (Fig. 145) is varied in making the resistance determination, the current in coil B will also vary, and if the current varies the amount of voltage induced in coil D will correspondingly vary.

Because of this condition it is necessary to have some adjustment in the power supply set by which the current in coil B can be maintained constant, without varying the frequency of the current. Later in this analysis a scheme will be shown for accomplishing this purpose.

The next point to be analyzed has to do with the constancy of current throughout the test circuit. Strange as it may seem, the current in coil L (Fig. 145) may be appreciably less than that in ammeter A, which presumably is measuring the current in the coil. This difference is generally due to voltage set up in the test circuit by electrostatic fields, and capacities to ground in the test circuit.

Suppose the test circuit is set up on a wooden table with iron framework, as indicated in Fig. 146. Suppose the coil B is going up and down (with respect to ground) in voltage, owing to the action of its exciting triode. Suppose an arrangement of apparatus as partly shown in Fig. 146, the thermocouple (or hot wire) ammeter A being over the grounded iron frame work of the table. The electrostatic induction from coil B into D (Fig. 145) will send charging current through ammeter A and this current may not go through any other part of the test circuit!

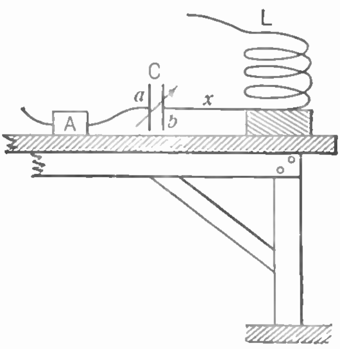


Fig. 146.

Sometimes this inequality of current throughout the test circuit occurs even when there is no appreciable electrostatic induction between coils B and D. This possibility also is indicated in Fig. 146. The ammeter A is close to ground (framework of the table) and the set of plates b (of condenser C) may be close to the framework. There may then be capacity current from ammeter A to the framework and back to plates b (or perhaps wire x) which do not go through the coil, L, and so vitiate the accuracy of the test results, which are based on the assumption that the same current flows all through the test circuit.

It is not difficult for the student to imagine other placements of the apparatus in the test circuit which will result in non-uniformity of current in the test circuit.

It is always best to use for condenser C one of the shielded type, in which the rotating plates of the condenser are connected to the shield. Coil L should then be connected to the insulated set of condenser plates, and so placed on the test table that its capacity to ground is as small as possible.

To insure the absence of any electrostatic inductive effects between the power supply circuit and the test circuit, several expedients are possible, and a simple test will show the adequacy of the precautions. If coil B is actually a part of the oscillating circuit of the triode, it should be connected in the oscillating circuit at a point as near ground potential (in so far as alternating voltage is concerned) as possible. Thus in Fig. 147, diagram a, coil B (which is always a small part of the total inductance in the oscillating circuit) is shown in the correct position in the oscillating circuit. Here it is at ground potential, so far as the radio frequency voltage is concerned. If coil B is connected as shown in diagram b, it will send out electric fields which may set up in the test circuit voltages much greater than those induced electromagnetically between coils B and D, upon which voltage the test circuit is assumed to depend for its current flow.

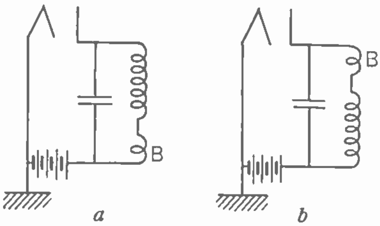


Fig. 147.

The whole scheme of resistance determination depends upon the assumption that, when the resistance introduced in R (Fig. 145) is sufficient to cut the current to 50 per cent of its original value (or some other convenient fraction), the introduced resistance is equal to (or some definite fraction of) the resistance of the circuit without the added resistance. But if the resistance of the test circuit itself changes, when the current is reduced to one-half its original value, the results obtained are meaningless and worthless, unless the amount by which the test circuit changes its resistance is known.

Now, of course, the temperature of the coil and condenser does not appreciably change, with variation of current, because even with maximum current the power loss in this test is so small that the whole circuit remains essentially at room temperature. However, the heater of the thermocouple ammeter does change temperature very much, when the current is halved, so it is necessary to know how the resistance of this heater varies with current. This phenomenon was investigated in Exp. 14, and unless it should happen that the heater used in this test is the same one as was measured in Exp. 14, it will be necessary to measure its current-resistance variation for this test.

This scheme of halving the current by adding resistance gives the resistance of the circuit, it will be noticed. It is then necessary to know the resistance of condenser, coil D, connections, etc., before the resistance of coil L can be determined. This determination is itself subject to considerable error, so it is well to have it small compared to the resistance of the coil being measured. The thermocouple heater should have a resistance reasonably small compared to that of the coil, condenser C should be of low loss, and with a known proportionality between series resistance and condenser setting, and the resistance of coil D and its connection should be small.

The total resistance of this circuit may be approximated then by measuring the resistance of it (with the condenser short-circuited) by continuous current, when the heater resistance is being measured in the bridge, as in Exp. 14. Of course, this determination is approximate only, but as the principal part of this resistance is the heater (and this has the same resistance for a.c. as for c.c.), and as the coil has a resistance large compared to the resistance of the rest of the circuit, the final error in coil resistance is not very large.

The equivalent series resistance of condenser C is obtained from the manufacturer's data, and law of resistance-capacity variation, if it has not been determined by other means. The manufacturer generally gives the data for a 1000-cycle frequency; its value for radio frequency is found by assuming that the series resistance, for a given setting, varies inversely with the frequency, and with (capacity)^2.

A condenser carefully designed and constructed of best material has a so-called " figure of merit " of about 6×10^{-14}. This figure of merit is given by the product $R\omega C^2$, in which R is the equivalent series resistance to represent the loss in the insulation blocks separating the two sets of plates.

If the series resistance, for a frequency of 1000 cycles, and capacity of one milli-microfarad, is 10 ohms, the figure of merit is

$$2\pi 1000 \times 10 \times (10^{-9})^2 \quad \text{or} \quad 6.28 \times 10^{-14}.$$

This is about as low a figure as is obtainable with ordinary materials. It connotes a reactance to resistance ratio for the condenser of about 1700.

If nothing is known about the resistance of this variable condenser C, it will be necessary either to neglect it altogether, or else to measure it. By making a " no-loss " condenser* of two heavy aluminum plates, one of them set in a copper shield, and the other hung by silk threads, outside of the electric field, the resistance of an unknown condenser

* See article by the author, *Proc. I.R.E.*, August, 1922.

can be reasonably well determined. A series resonant circuit like that of Fig. 145 is arranged so that the no-loss condenser can be substituted for the other. It will be necessary to put a resistance box in series with the no-loss condenser, to give the same current in the circuit when it is substituted for the unknown, and this amount of resistance is the resistance of the unknown condenser.

As the resistance will be very small, great care must be exercised to eliminate errors due to mistuning, etc., when one condenser is substituted for the other, if a reliable measurement is to be made.

To keep the errors in this experiment small it is well to have the power-generating triode and its circuits in a copper-lined box. The coil *D* is placed inside the box, and arranged for conveniently changing its coupling with coil *B*. An electrostatic shield, grounded, should surround coil *B*. This is conveniently made of copper strips or wires arranged as a cage around coil *B*, the strips being parallel to each other and connected together along only one edge, as indicated in Fig. 148. This scheme permits the changing magnetic field of coil *B* to reach through and induce the desired voltage in coil *D*, but prevents any electric field from changing the potential of coil *D* with respect to ground.

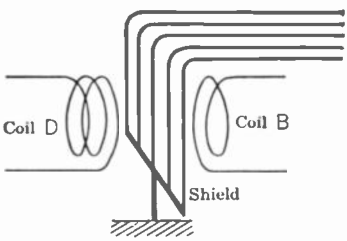


FIG. 148.

The rest of the test circuit (*L*, *C*, *R*, and *A* of Fig. 145) may well be in another copper-lined box, with windows in the cover for observation and to permit changes in *C* and *R*. The box should be large enough that coil *L* does not come close to the copper. The coil should be held up by suitable pegs of dry wood.

The power from coil *D* should be supplied to the test circuit through a small *DPDT* switch arranged as a reversing switch. If there is no electrostatic induction getting into the test circuit from the power circuit the reading of the ammeter *A* will be the same whichever way the switch is thrown. If different readings are observed, suitable changes in the arrangement of the circuits must be made until the difference is eliminated.

Measure the resistance and inductance of two or three typical radio coils, throughout the broadcast band of frequencies.

With three single-layer solenoids wound with the same size and turns of wire, with different spacings between wires, get curves of resistance *vs.* frequency, to find effect of wire spacing.

With a small two-layer solenoid (not bank wound) of about 200 microhenrys inductance, get curves of resistance and inductance *vs.* frequency. With this coil the resistance and inductance both increase with frequency, especially as the natural period of the coil is approached. With the single-layer solenoids the inductance will be found almost independent of the frequency.

Measure the resistance (for a suitable frequency range) of a portion of a coil, say between taps including about one-fifth of the turns of a single layer solenoid of one or two millihenrys inductance.

EXPERIMENT 30

Object.—To measure the resistance, inductance, and capacity of an antenna at various frequencies, both above and below its natural frequency.

Analysis.—The simplest type of antenna to conceive is of course the simple vertical wire; practically, however, it is impossible to suspend such a wire without having poles of some sort, which of course raise the "ground" up around the antenna, and by the currents flowing in them, change the measured characteristics to a great extent. So we will consider first the long horizontal antenna which can be hung up in the air without closely adjacent poles. It must of course have a vertical portion, *A–B* in Fig. 149, called the "down lead"; we shall at first neglect the down lead and consider only the horizontal portion *B–C*. For low-frequency antennas this portion may be 10 or 20 times as long as the portion *A–B*, but for short-wave antennas the part *A–B* is frequently the major part of the antenna. If we imagine a 60-cycle alternator connected between point *A* and the underlying ground, a very small charging current will flow into the antenna. For an antenna of $0.001 \mu f$ capacity and an alternator of 100 volts at 60 cycles the current will be about 40 microamperes.

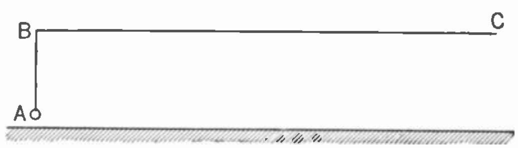


FIG. 149.

The potential of the antenna with respect to ground (which is considered as zero potential) will be everywhere the same, at a given instant. The current of 40 microamperes occurs only at the base of the antenna; a little reflection shows that it decreases linearly as we approach the open end of the antenna, evidently being zero at the point *C*.

The capacity of the antenna, determined from the relation $I = \omega CE$, will be equal to the capacity per unit length, multiplied by the length of the antenna. This is the same capacity as would be determined by charging the antenna from a battery of known voltage and discharging it through a ballistic galvanometer. It is also the same capacity as would be obtained by using the relation,

$$\text{energy stored in the electric field around the antenna} = \frac{CV^2}{2},$$

where V is the potential of the antenna with respect to ground.

The inductance of the antenna might be determined by grounding the distant end, and, from the alternator voltage and current, calculating the impedance. Allowing for the resistance of the antenna the reactance could be calculated and from this the self-induction. The self-induction thus obtained would be equal to the self-induction of the wire per unit length, multiplied by the length of the antenna.

In this test the current throughout the length of the antenna would be the same, at any instant; this is then entirely different from the condition existing when the distant end of the antenna is insulated from earth. Here the current decreases linearly as the distant end is approached, so that quite evidently the energy stored in the magnetic field around the antenna, for a given current at A (Fig. 149) is much different than for the case when the distant end is grounded. If we define the self-induction from the viewpoint that energy stored in the magnetic field is given by $LI^2/2$, where I is the current at A, then the inductance determined from the open circuit test will be only one-third as much as for the case when the distant end of the antenna is grounded.

Now as the frequency of the impressed voltage is increased, both current and voltage change their distribution along the antenna. Of course the current at the distant end of the antenna must always have zero value whatever the frequency, but this is about the only condition that stays fixed.

Up to a certain critical frequency (which gives quarter wave length distribution, and which we further discuss below) the antenna acts towards the impressed voltage as a condenser; it draws a leading current from its power supply. With constant impressed voltage, and increasing frequency, the current increases in magnitude slowly at first and then much more rapidly, reaching a maximum value at a definite frequency. If this frequency is exceeded the current rapidly diminishes.

Now suitable measurements show that at the frequency showing maximum current, the antenna no longer acts like a condenser, but like a resistance of reasonably low value, from 5 to about 50 ohms. The power factor of the antenna at this frequency is unity; the reactance is zero. For this condition the distribution of voltage and current in the antenna is as shown in Fig. 150; both current and voltage are distributed along the antenna in a nearly sinusoidal curve. The current, starting at the base of the antenna with a value of I_0 diminishes on a sine curve to its zero value at the extreme end of the antenna. The voltage of the antenna with respect to ground starts at the value E_g (the generator voltage) at the base and rises on a sine curve to a value of E_t at the open end of the antenna. This value E_t may be from 10 to 50 times as large as E_g, the voltage impressed on the antenna by the power supply.

Now the velocity of travel of an electric disturbance along a straight wire is practically the velocity of light, V. The relation between frequency and wave length of any wave motion is given by the general formula $V = f\lambda$, in which f is the frequency and λ is the wave length. As $V = 3 \times 10^8$ meters per second it follows that

$$\lambda_{\text{meters}} = \frac{3 \times 10^8}{f}. \qquad (69)$$

The frequency of impressed voltage which gives the distribution of Fig. 150 must then be such that its wave length is four times as long as the antenna. Thus if the down lead in Fig. 150 is 10 meters and the horizontal length is 30 meters, the wave length to set up the condition in Fig. 150 is $4 \times (10 \text{ plus } 30) = 160$ meters. And the frequency is

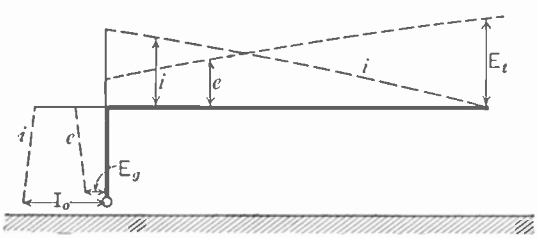


FIG. 150.

$$3 \times 10^8/160 = 1{,}875{,}000 \text{ cycles} = 1{,}875 \text{ kilocycles (kc.)} = 1.875 \text{ megacycles.}$$

If the antenna of Fig. 150 is suspended in a vertical position (say by a balloon) and it is grounded through a low-resistance connection, the resistance for the quarter wave length oscillation will be about 37 ohms; it will have this value no matter what the length of the antenna may be, if only the impressed frequency is such as to set up the quarter wave length oscillation. This resistance is the *radiation resistance* of the antenna; a well-grounded antenna, suspended vertically from a balloon, will have but little other resistance, at its quarter wave length oscillation. We say " suspended from a balloon " because when hung between steel towers (as is the case with the average antenna) the radiation resistance may be only a small fraction of this value; currents induced in the towers and other parts of the suspension structure may reduce the radiation resistance of the antenna to a very small fraction of its theoretical value.

Losses in heating the wires of the antenna itself, losses in insulators, losses in the ground under the antenna, etc., all act to increase the actual resistance of the antenna, especially for the lower frequencies. In Fig. 151 is shown a conventional antenna resistance curve, showing a resistance of about 35 ohms at the quarter wave length excitation, λ_o; this diminishes to about 10 ohms at about half this frequency, and for still lower frequencies it rises on a nearly straight line.

The radiation resistance, varying inversely with the (wave length)^2, is shown by the dotted curve, falling rapidly for the longer wave lengths. It is to be remembered that the power radiated from the antenna is equal to this resistance multiplied by the square of the antenna current. Whatever other resistance the antenna may have (in addition to the radiation resistance) represents loss of power only; it is of no service in carrying on radio communication, but results in heating of the ground and other material in and around the antenna.

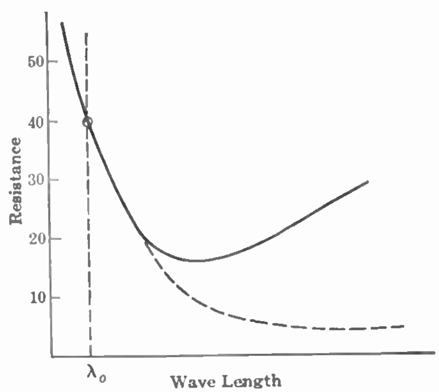


FIG. 151.

It is practically necessary to tune the antenna circuit to the frequency of its power supply; otherwise practically no energy will flow into the antenna from the power source. For frequencies of power supply lower than the quarter wave length frequency, it is necessary to insert an inductance between the antenna and ground, in order to establish resonance with the power supply. A coil so used is generally called a loading coil.

If it is desired to supply the antenna with power of frequency higher than the quarter wave length frequency it is necessary to connect a condenser between the antenna and ground. Such a condenser is evidently *in series* with the capacity of the antenna and so results in a total capacity in the antenna circuit less than the antenna capacity. Such a condenser, inserted in series with the antenna to tune it for frequencies above its natural frequency, is called a "shortening condenser," as it results in a shorter wave length radiation.

If the true inductance, L_0 (for uniform current distribution) and true capacity, C_0 (for uniform potential distribution) are measured by any means it will be found that they do not serve to calculate the natural frequency (quarter wave length) from the formula

$$f = \frac{1}{2\pi\sqrt{L_0C_0}}.$$

It will be found that the frequency so calculated is too low; it must be multiplied by $\left(\frac{\pi}{2}\right)^2$ to give the quarter wave length frequency.

As both the L and C of the antenna, for quarter wave length oscillation, are determined by a sinusoidal distribution of either current or voltage, as L_0 and C_0 are determined with uniform distribution of current and voltage, it is reasonable to assume that both L_0 and C_0 have been reduced in the same degree by the redistribution of current and voltage. Thus for the quarter wave length oscillation $L_a = (2/\pi)L_0$ and $C_a = \frac{2}{\pi}C_0$; using these values L_a and C_a in the resonance formula $f = \frac{1}{2\pi\sqrt{L_aC_a}}$, does give the natural frequency.

The above theory deals only with antennas of uniform structure, but very few antennas are so constructed. Most antennas are of T form, and the top part of the structure generally consists of several wires in parallel, whereas the down lead is a single wire. The above simple theory can be applied only approximately to this structure, because the inductance and capacity per unit length vary at different parts of the antenna. However, such an antenna is affected by loading coils and shortening condensers in the same manner as the simple antenna.

As mentioned in Exp. 17, it is customary to measure the capacity of an antenna for the uniform potential distribution condition. This is done by measuring the resonant frequency of the antenna when a large loading coil has been added. From the measured frequency and known inductance (antenna inductance neglected) the capacity of the antenna is calculated. From this value of capacity and the measured quarter wave length frequency the inductance of the antenna is calculated. Correction to this calculated value of capacity can then be made if desired, as shown in Exp. 17.

In this experiment the antenna constants are to be determined by comparison with a dummy antenna, the circuit being arranged as in Fig. 152. A variable frequency power supply, exciting coil M, induces a current in the antenna by the coupling between M and the loading coil L. The magnitude of the current is read by the thermocouple ammeter A.

For a given value of loading coil L, the frequency of the power supply is varied until ammeter A indicates that resonance has been attained. The double-throw switch S is then thrown so as to include the variable low loss, calibrated condenser C, and resistance box R, in series with the loading coil L and ammeter A. Condenser C is varied until resonance is established with the power supply, and then the resistance R is varied until ammeter A reads the same as it did in the antenna circuit. After checking the resonance conditions, and equality of ammeter readings, by means of S, the readings of C and R may be taken as the capacity and resistance of the antenna.

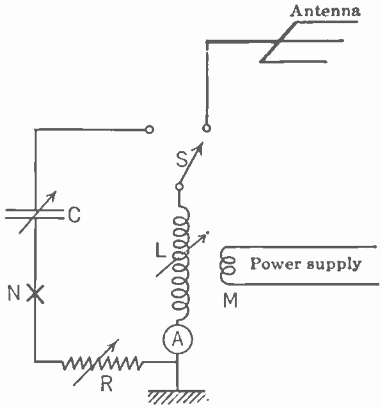


FIG. 152.

This operation is repeated for a whole series of loading coils, preferably constructed for this purpose, of low-resistance radio cable. The lower the resistance of the loading coils, the more accurately can the antenna resistance be determined. The resistance of the thermocouple heater should also be as low as feasible.

In this determination the inductance of the antenna itself has been neglected. If it is desired, this can be allowed for by introducing into the dummy antenna circuit, at N, a small coil of the same inductance as the antenna. Its resistance must be known, by other tests, throughout the frequency range to be used in this test.

After making measurements with all values of loading coil called for in this experiment, the measurements should be carried out for two or three values of shortening condenser, substituted in place of the loading coil L. These should preferably be small fixed condensers of mica, so that their series resistance is small. Suitable values of capacity are same capacity as the antenna, twice as much, and five times as much.

In this test the loading coil L cannot be completely removed. A small coil must be left in circuit, to serve for coupling the antenna circuit to the power supply.

If the high-frequency resistance of the various loading coils is known, the resistance values of the antenna can be checked by the added resistance method. No dummy antenna circuit is required, but the resistance R is inserted directly in series with the antenna. After establishing resonance with the power supply, the value of R being zero, sufficient resistance is put in R to reduce the current to one-half its value. The added resistance required to bring this about is just equal to the resistance of the antenna circuit, so that if the resistances of the loading coil and ammeter A are known, the resistance of the antenna is at once determined. It is generally necessary to retune the antenna circuit, by changing the impressed frequency a slight amount, after the resistance R has been added. This is because the resistance R itself has some inductance which may be sufficient to mistune the antenna to an extent sufficient to affect the ammeter reading.

If the antenna is located an appreciable fraction of a wave length from other conducting structures the resistance-frequency curve will probably be smooth, but if it is located on a steel framework building or is near metal roofs, poles, etc., there will probably be found many irregularities in the resistance curve. An antenna which shows a resistance curve rising rapidly with increasing wave length generally has a poor ground connection, or has much defective dielectric in place to be affected by the electric field of the antenna. Cases have been reported which indicate that a tree in the immediate neighborhood may possibly double the resistance of an antenna, at the lower frequencies.

Plot suitable curves to show how the antenna constants change with wave length. Show by dotted line the estimated radiation resistance of the antenna.

EXPERIMENT 31

Object.—Study of the action of a triode in an oscillating receiver circuit. Interpretation of noises heard in the telephone receiver. Action of too much coupling, when using grid condenser and leak. Heterodyne reception of continuous wave signals.

Analysis.—When a triode is used to generate alternating current power in circuits requiring watts of power, such as those studied in Exp. 27, c.c. meters are used in grid and plate circuits, and hot wire meters serve to measure the amount of high-frequency current flowing in the oscillating circuit. In circuits of this type it is comparatively easy to grasp quantitatively the interdependence of the various circuit adjustments.

In a certain type of radio receiving set it is necessary to arrange the circuit to generate alternating currents of adjustable frequency, and the magnitude of the alternating current set up is so small that no ordinary meter would measure it, even if the receiving set were equipped with such a meter. Furthermore, no meters are furnished to measure either plate or grid current; the only instrument available to tell the operator what is happening in the circuit is the telephone receiver in the plate circuit of the triode. This, of course, is not a quantitative measuring device; in combination with the ear it is only a very rough meter of the amount of energy supplied to it. (See the results obtained in Exp. 23, p. 179, on ability of ear to judge intensity of sound.)

In this experiment the student is to set up one of these oscillating circuits, having, however, a milliammeter in the plate circuit in addition to the telephone receiver; by observing the indications of the plate current meter in conjunction with the sounds heard in the phone, the significance of the phone sounds is readily grasped.

The signals of a continuous wave transmitter, if put through a crystal rectifier receiver, or a non-oscillating triode detector, would be inaudible except for faint undecipherable clicks. To produce ordinary musical dashes and dots, it is necessary to have the incoming signals coexist in the receiver circuit with another high-frequency current of constant amplitude, the frequency of which differs from that of the incoming signal by a few hundred or a thousand cycles per second. These two coexisting high-frequency currents of different frequencies will produce "beats," that is, pulsations in amplitude of the actual complex current existing in the circuit, and these pulsations in amplitude will have a frequency equal to the difference of the signal frequency and the locally generated frequency.

A varying amplitude, high-frequency current, passed through the ordinary detector circuit, produces a current in the phones of frequency the same as the amplitude variations. This then is the frequency of the note the listening operator hears. It is under his control because he can adjust the frequency of the locally generated oscillations. The circuit generally used employs a tuned circuit in the grid of the triode, coupled magnetically to a coil in the plate circuit, as shown in Fig. 153. The coil L_3 which supplies plate circuit energy to the grid circuit is generally called the "tickler" coil; it is of comparatively few turns, arranged for adjustable coupling with L_2. With the correct relative polarities of coils L_3 and L_2, as the coupling is increased it will be found that oscillations are set up in the L_2–C_2 circuit when M exceeds a certain value, the value depending upon plate voltage, resistance of the L_2–C_2 circuit, grid bias, amount of coupling between L_2 and L_1, value of L_2, filament current, etc.

Let us first suppose there is no coupling between L_2 and L_1, so that the disturbing effect of the antenna is absent.

When the coupling of L_3 to L_2 is increased past the critical value, oscillations begin, and we know that under ordinary conditions these make the average value of the plate current increase; this effect was investigated in Exp. 22, in which arrangement the increase in plate current was used as a measure of the amplitude of the alternating voltage impressed on the grid circuit.

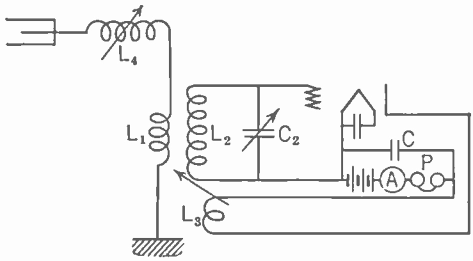


Fig. 153.

Now when oscillations are set up in such a circuit as that of Fig. 153, this increase in average plate current must flow through the telephone receivers, thus pulling down the diafram, or letting it pull away from the pole pieces, according to the direction of current through its windings.

Such a motion of the diafram, if it takes place quickly enough, will give a click, the sharpness and distinctness of the click depending upon the rapidity with which the plate current changes. By having a milliammeter in series with the phones, and listening carefully, a faint click will be heard as M is increased, and this click will be heard just as the milliammeter in series with the phones shows the increase in plate current.

The increase in M must be carried out with sufficient speed or else the oscillations are set up so gradually that the plate current increases too slowly to produce an audible sound in the phones. It will be noticed in this test that the click becomes softer and more indistinct as the increase in M is brought about more slowly. It will be noticed from the plate circuit ammeter that the sharpness of the click depends entirely on the speed with which the plate current increases, that is, upon the speed with which M is increased.

When M, being diminished, is decreased past the critical value, it may be that another click is heard, if M is decreased rapidly enough.

It is possible to make the plate current decrease when oscillations start, by underexciting the filament, that is, using less than normal filament current. If the characteristic I_p–E_g curve of the triode has the form shown in Fig. 154, the plate current will decrease when oscillations start. The upper, level, part of the curve is due to lack of emission, caused by the filament current being too low. Evidently when the grid potential oscillates about point A (Fig. 154), the decrease of plate current when the grid goes negative is much greater than the increase when the grid goes positive. So the average plate current decreases.

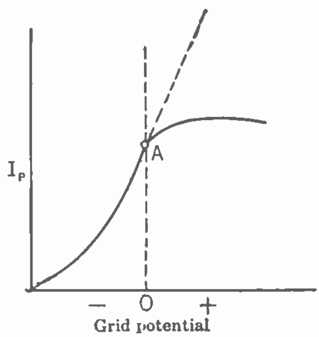


Fig. 154.

However, if the filament current is increased to give sufficient emission, the I_p–E_g curve follows the dotted line; the increase of plate current, with alternating voltage on the grid, is greater than the decrease, so the average plate current increases.

Frequently the arrangement of Fig. 153 is modified by the use of a condenser, shunted by a leak resistance, in series with the grid. In this case the plate current always decreases when an alternating voltage is impressed on the grid, so that the starting of oscillations for this circuit is always accompanied by a *decrease* in plate current. This effect also was investigated in Exp. 22.

As the average oscillating receiver circuit has no ammeter in the plate circuit, the operator must tell by the sound in his phones whether or not the circuit is in a state of oscillation. As already analyzed, there is a slight click when oscillations start or stop, provided the effect is rapid enough. However, there is a much more reliable and simple test to detect oscillations.

If the finger (preferably moistened) is touched to the grid, the oscillations are practically always stopped, if there are any. The slight coupling between plate and grid circuits is not sufficient to supply the loss due to high-frequency currents flowing in and out of the body, and so the oscillations are stopped. With tighter coupling it may be necessary to actually connect the grid to the filament, by means of finger and thumb, to stop the oscillations. This is done by touching the filament terminal with the thumb and the grid with the finger. As contact with the grid is made a sharp click is heard, and an equally sharp one is heard when the finger is removed. There is only one condition under which this finger test does not always give reliable indications and this is shown with the help of Fig. 155. The grid leak may connect around the condenser C, or it may connect from the grid to the positive filament terminal; this is the preferable connection for the average triode.

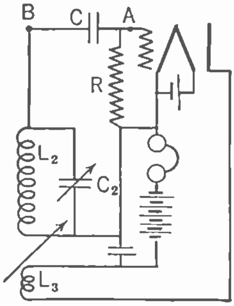


FIG. 155.

The grid is not now at the same potential as the filament, even when there are no oscillations, so that when the thumb is in contact with the filament and the finger is touched to the grid, the potential of the grid changes and a corresponding change in plate current occurs. Thus a click is heard in the phones, and the triode is not oscillating.

If, however, the finger is touched at point B (Fig. 155) instead of point A, the test for oscillations is a reliable one. Point B is evidently at the same potential as the filament because it connects to the filament through the negligible resistance of coil L. Hence, when there are no oscillations present, making finger contact with point B gives no click in the phones, but a very sharp click is heard if there are oscillations present. This is because the oscillations are stopped when the finger is touched to point B; stopping the oscillations changes the average grid potential, and so the average plate current.

The finger test for oscillations is therefore reliable if the thumb is touched solidly to the filament and the finger is touched solidly to that end of coil L which is next to the grid. *Presence of a click when contact is made or broken means oscillations are present, and of course vice versa.*

The use of grid condenser and leak is advisable because of the much greater sensitivity of the triode as a detector when a condenser is used. However, certain peculiar effects may be observed, with this connection, which we will now analyze.

If the coupling between the tuned circuit and tickler coil is made too tight, a shrill note may be heard in phones, with certain values of condenser and leak resistance. With other values a low musical note may be heard, or even a series of clicks. The note, or frequency of clicks, is generally lowered if the tickler coupling is increased, and generally raised if the capacity of the condenser in the tuned circuit is increased.

By increasing the capacity of grid condenser, and increasing the resistance of the leak, other conditions being left the same, the frequency of the note, or of the clicks, may be lowered until the click occurs perhaps at the rate of *once a minute or slower.* To obtain this frequency the grid condenser must be a microfarad or more, and the leak resistance many megohms. Also the triode itself must have a high resistance between grid and cathode, that is, there must be very little ionized gas in the tube.

The phenomenon mentioned above is caused by the triode generating oscillations (of about the natural frequency of the tuned circuit) in groups, and the frequency of these groups is the note heard in the phones.

With the tight coupling assumed when oscillations start in the L_2–C_2 circuit (Fig. 155) they become so violent that a very large negative charge builds up on the grid. The resulting high negative potential on the grid completely shuts off the plate current, causing it to fall to zero. But this leaves the tuned grid circuit with no power supply; it must be remembered that the power to maintain the oscillations in the grid circuit comes from the fluctuating plate current.

The oscillations in the grid circuit cease almost as soon as the plate current is shut off, because of the decrement of the circuit. Thus if the decrement is 0.03, a reasonable value, the oscillations will die down to 1 per cent of their original amplitude in $\frac{4.6 + 0.03}{0.03}$ cycles, or 154 cycles. For a circuit having the right L_2 and C_2 for tuning broadcast frequency currents, this represents only about 0.0002 second. Thus we may say the oscillations die out practically as soon as the plate current drops to zero.

Now the charge on the grid condenser must practically all leak off before the plate current can again increase to a value sufficient to start oscillations. The time taken for the charge to leak off is determined by the time constant of the circuit, it taking a time equal to CR seconds for 63 per cent of the charge to leak off.

If the grid builds up to a negative potential of 50 volts, and not more than 5 volts bias on the grid can be used without decreasing the plate current to such an extent as to cause cessation of oscillations, then 90 per cent of the grid charge must leak off before oscillations again start. This requires a time about 2 CR seconds. That is, if $C = 1\mu f$ and $R = 1$ megohm, two seconds must elapse before the oscillations again start, after being stopped.

This effect can be studied very easily by means of the plate circuit ammeter. Simultaneously with a click in the phones, this ammeter reading drops to zero; presently a gradual increase in the plate current is seen (the charge is leaking off the grid condenser), and when it reaches a certain critical value another click is heard in the phones and the ammeter again drops to zero.

With the ordinary grid condenser of $250\mu\mu f$ and leak of 2 megohms the time constant is only 0.0005 second. According to the analysis given above, the plate current would rise and then fall to zero about 1000 times a second, thus giving 1000 clicks in the phones each second. This would sound like a pure note of 1000 vibrations per second.

Now if the coupling is increased, other conditions being left the same, the oscillations start more violently than they did before, the grid is forced to even greater negative potentials when the oscillations start, and thus, owing to the greater charge gained by the grid, more time is required before the charge can leak off. This results in a lowering of the note heard in the phones.

If the value of capacity in the L_2–C_2 circuit is increased, the intensity of oscillations set up by a given coupling is diminished, so that the charge built up on the grid condenser is not as great. This would tend to give a higher frequency of clicks in the phones. However, oscillations are not established as easily with a larger value of C_2 so the plate current must have a higher value before oscillations are re-established. This effect tends to give a lower frequency of clicks in the phones.

The first action generally has the predominant effect, because it will be found that the note in the phones generally increases in frequency as the capacity of the tuning condenser is increased.

The coupling between the tickler coil and tuned grid circuit must be limited to a value only slightly above its critical value (that is, the value required to start oscillations), if this periodic starting and stopping of oscillations is to be eliminated. This necessitates that the tickler coupling be increased, as the tuning condenser is increased in capacity. If sufficient coupling is used to properly establish oscillations with all of the tuning condenser in circuit, it will generally be found that the circuit will squeal (start and stop oscillations) if the tuning condenser is decreased to its minimum value, the coupling being left the same. For best results, coupling must be diminished as tuning condenser is decreased.

We will now consider the effect of the antenna circuit; a careful study of its action will show the applicability, to radio circuits, of the results obtained in Exp. 5. It was shown there that a tuned circuit, coupled to a coil, always increased the effective resistance of the coil, and that it might either increase or decrease the reactance of the coil according as its condensive or inductive reactance predominated.

Suppose now the coupling between L_2 and L_3 (Fig. 153) has been increased just enough to produce oscillations in the L_2–C_2 circuit, as shown by the finger test. If, now, with the antenna circuit coupled to the L_2–C_2 circuit, the antenna circuit is tuned to the L_2–C_2 circuit, it will be noticed that the oscillations in the L_2–C_2 circuit are suppressed. The finger test will show that there is a range of adjustment of the tuning of the antenna circuit within which the triode does not oscillate; this range increases as the coupling between the antenna and L_2–C_2 circuit is increased. If the tuning of the antenna circuit is changed with sufficient rapidity the phones will give a double click as the tuning passes through the critical range.

It will be found that oscillations in the L_2–C_2 circuit can be restored, even when the antenna is tuned to this circuit, by increasing the tickler coupling to a sufficient extent. The tighter the antenna is coupled to the L_2–C_2 circuit, the more must the tickler coupling be increased to produce oscillations.

The amount of tickler coupling required to produce oscillations depends upon the resistance of the L_2–C_2 circuit; the greater the resistance, the more must the coupling be increased. Now the coupled antenna circuit greatly increases the effective resistance of the L_2–C_2 circuit, when the antenna circuit is tuned to the frequency of the current in this latter circuit. It is because of this resistance, introduced into the L_2–C_2 circuit by the tuned antenna, that the L_2–C_2 circuit oscillations are suppressed, unless the coupling of the tickler coil is increased sufficiently to overcome this effect of the additional resistance.

This cessation of oscillations due to the resistance introduced into the L_2–C_2 circuit by another circuit, coupled and tuned to it, is useful in measuring the frequency of the oscillations. If the extra circuit is a wave meter the cessation of oscillations is an indication that the wave meter is tuned to the frequency of these oscillations; by sufficiently diminishing the coupling between the L_2–C_2 circuit and the wave meter, the range of wave meter adjustment throughout which the oscillations are suppressed becomes narrower, so that a quite accurate frequency determination of the oscillations can be made. Some such scheme as this for measuring frequency is necessary in oscillating receiving circuits; the amount of power generated in the oscillating circuit is not sufficient to operate the hot wire meter by which the wave meter is ordinarily tuned, so that it is only by this indirect indication, in the circuit being measured, that the wave meter is tuned to it.

The effect of the antenna circuit on the reactance of the L_2–C_2 circuit can be nicely determined, in a qualitative manner, by listening to the beat note between the oscillating receiver, and a continuous wave transmitter. The latter can well be a small transmitting triode circuit set up right in the laboratory, and left oscillating.

When the tuning of the receiving circuit (this being adjusted to oscillate) is slowly changed, through a value which tunes it to the transmitter, the following effect will be noticed. A shrill whistling note, at the upper limit of audibility, is first noticed, and as the tuning condenser of the receiver is changed this note comes gradually down the whole audible scale, passes below audibility, appears again as a low note, and then gradually ascends and disappears again through the upper limit of audibility.

Let us suppose an ordinary SLC condenser is used and that it is set at 50 (on a 0 to 100 scale) to generate a frequency of 1000 kc., and that the laboratory transmitter is generating 1000 kc. An ordinary phone is extremely inefficient at the upper limit of audibility, so we may regard 10,000 vibrations a second as high as the telephone will give.

If the tuning condenser of the receiver is set at 45 the frequency generated will be about 1050 kc.; the beat note of 50 kc. will of course be inaudible. When the condenser is set at 49 the frequency generated is 1010 kc., so the beat note becomes audible. A variation from 49 to 50 on the tuning condenser changes the beat note from 10 kc. to zero, so this slight change in setting of the tuning condenser, from 49 to 50, is sufficient to send this heterodyne note through the whole audible scale.

If now the condenser is changed from 50 to 51 the note, starting from zero, goes up to a frequency of 10,000 vibrations per second, and further increase in capacity setting sends it through the upper limit of audibility.

Unless the circuit is properly grounded it may be that the note changes by thousands of vibrations per second, as the hand is approached to tune the circuit. This is due to the capacity of the body affecting the frequency of the oscillations generated in the receiver.

If the transmitter is close to the receiver a peculiar effect will be noticed. As the note heard in the phones approaches the lower limit of audibility it disappears. Examination of the condenser setting may show that the difference of transmitter and receiver frequencies should be, say, 500, yet no note is heard. It is naturally supposed that the receiver has stopped oscillating, for some cause or other, but the finger test shows that the circuit is oscillating. By observing this phenomenon with the transmitter at different distances from the receiver it will be found that the *receiver tends to synchronize itself with the transmitter*, and that the ability of the transmitter to pull the receiver into synchronism with itself increases as the distance between them is decreased. For a certain distance between transmitter and receiver it will be found that a certain minimum frequency beat note can be heard. As the coupling between transmitter and receiver is increased, the pitch of this " lowest audible note " increases, so that if they are close together, for example, they may pull into synchronism when their natural frequencies (each independent of the other) are possibly 1000 cycles apart. This means that the lowest beat note audible has a pitch of 1000 vibrations per second.

Besides the normal heterodyne note which has a pitch equal to the difference of the two frequencies, many other beat notes can be heard for various adjustments of transmitter and receiver. The transmitter necessarily generates a number of harmonic currents, in addition to its fundamental. So if its normal frequency is 1000 kc., as assumed above, it is also sending out 2000 kc., 3000 kc., etc. We supposed above that the receiver condenser, when set on 50, gave its circuit a frequency of 1000 kc. If then it is set on about 12 it will generate 2000 kc., and so a beat note will be heard between the receiver frequency and the second harmonic of the transmitter. With its condenser set on about 4, the receiver circuit generates 3000 kc., so a beat note is heard from the third harmonic of the transmitter. In this way a great many harmonics of the transmitter may be picked out.

It also may occur that the receiver circuit is set to generate 500 kc. and that it generates as well a second harmonic of 1000 kc. Then a beat note may be heard between the fundamental of the transmitter and the second harmonic of the receiver.

It will be found that the relation between condenser setting of the receiver, and the frequency it generates, is not quite in accordance with the law that frequency varies inversely as the square root of the capacity. This is due to the fact that the circuit itself has capacity between its various parts in addition to the condenser capacity, and, of course, as the condenser setting is changed the stray capacity is not altered.

The effect of this stray capacity is especially noticeable when the condenser has comparatively low capacity, that is, at the lower divisions of its scale.

When listening to the beat note phenomenon, the effect of the antenna tuning on the reactance of the coil L of the receiver circuit can easily be observed.

Suppose the transmitter is set to give a frequency of 998 kc. and the receiver is set to give 1000 kc., with no coupling between it and the antenna. The beat note heard has a pitch of 2000 vibrations.

Now we will suppose that the antenna is coupled to the coil of the oscillating circuit, with too small an inductance for tuning it to the frequency of the L_2–C_2 circuit. The capacity reaction of the antenna then exceeds the inductive reaction, and the current in the antenna circuit is a leading one. The results of Exp. 5 show that under such conditions the antenna circuit will increase the apparent inductance of L_2. This increase is small until the antenna is nearly tuned to the impressed frequency (that of the L_2–C_2 circuit, in this case) and then the effect increases rapidly and as suddenly decreases to zero, when the antenna is exactly tuned to the L_2–C_2 circuit. If the antenna inductance is further increased, the preponderant reaction in this circuit becomes inductive, the current lags, and the apparent inductance of L_2 is decreased by the action of the antenna circuit. This demagnetizing action exerted on L_2 is large when the antenna is not far from the tuned condition, then falls off and becomes negligible.

Both of these effects, the magnetizing and demagnetizing action of the antenna circuit, increase with the degree of coupling between antenna and the L_2–C_2 circuit.

As stated above, with too little inductance in the antenna the apparent inductance of L_2 is increased. This will decrease the frequency of oscillation of the L_2–C_2 circuit. Hence instead of generating 1000 kc., as is the case with no antenna coupling, it generates, say, 999 kc. This makes the beat frequency 1 kc. instead of 2 kc., as it was with no antenna reaction. The antenna thus decreases the note heard in the phones.

As the inductance in the antenna is increased to the proper value for tuning, the beat note returns to its 2-kc. value. As the inductance is further increased, the L_2–C_2 circuit oscillates at perhaps 1001 kc., owing to the decreased effective value of L_2. The beat note thus comes up to 3 kc., and with further increase in antenna inductance the reaction back into the L_2–C_2 circuit becomes negligible, and so the beat note returns to 2 kc.

If we had assumed the L_2–C_2 circuit was oscillating at 996 kc., the beat note would have been 2 kc., as it was before. But in this case the antenna with too little inductance would have raised the beat note to 3 kc., and with too much inductance in the antenna for tuning the beat note would have dropped to 1 kc.

Of course it will be appreciated that the concrete values given above are illustrative only. The change in pitch of the beat note may be greater or less than the amounts stated, depending upon the circuit constants. But qualitatively the experimental results obtained in this test will be the same as those predicted by the above analysis.

With a circuit arranged as in Fig. 155 try the effect of varying the tickler coupling, having it connected in one way, and then with its connections (or relative polarity of coupling) reversed. When critcial coupling is reached, note any sound in the phones and indications of the milliameter in the plate circuit. Change the coupling rapidly, to see whether the click in the phones becomes more or less distinct.

Try the finger test, noting sound in phones and ammeter reading.

Try same tests with a circuit having grid condenser and leak. Try condensers of various capacities and various leaks. Note and explain peculiar effects when using too tight a coupling. Use largest value of capacity available and no leak (other than that provided by the tube itself) to see how slowly the periodic clicks may be made to occur.

Try effect of coupling another circuit (to resemble an antenna) to the coil in the tuned circuit. With this second circuit tuned to the L_2–C_2 circuit note how tickler coupling must be increased (to maintain oscillations) as antenna circuit coupling is increased.

Start transmitter circuit and study heterodyne method of signal detection. Note effect of harmonics discussed in the analysis. Note synchronizing effects and effect of coupling between transmitter and receiver.

Study effect of antenna on reactance of the L_2 coil by analyzing the change in pitch of the beat note heard in the phones, as the tuning of the antenna circuit is varied. Note carefully how the pitch changes for both conditions, oscillating circuit L_2–C_2 tuned above the transmitter frequency and also below the transmitter frequency. (If a coupler L_2–L_3 is used in which one coil rotates closely inside the other, it may be found that even with the coils at right angles [zero mutual induction] oscillations are set up. This is due to the capacity present between the two coils, acting as two plates of a condenser. This capacity between the plate and grid circuits may give sufficient coupling to sustain oscillations even if there is no magnetic coupling. It is well to have the two coils separated from each other by as much distance as compatible with the requisite magnetic coupling. Thus the inner coil of a rotating coupler should be of considerably smaller diameter than the outer.)

EXPERIMENT 32

Object.—Quantitative study of conditions for setting up oscillations in typical triode circuits.

Analysis.—In previous tests the conditions for setting up oscillations have been qualitatively studied for two or three cases; it is the object of this test to examine the conditions for producing oscillations more accurately, with the idea of checking the mathematically deduced requirements for making certain typical circuits oscillate. The first circuit to be examined has the tuned circuit in the plate, the oscillations being produced by exciting the grid through the mutual induction of L_1 and L_2 (Fig. 156). Such a circuit as this is examined mathematically by supposing that some disturbance sets up a small current in the L_1–C_1 circuit. Such a current will evidently be oscillatory (unless R_1 is excessively high), and will die down according to an exponential law.

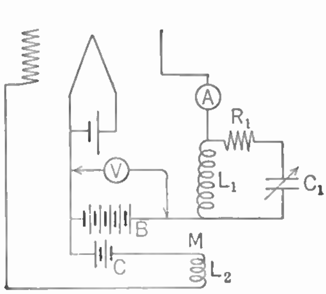


FIG. 156.

The oscillatory current in the L_1–C_1 circuit will, however, induce a voltage in L_2, thus making the grid go up and down in potential, as the current alternates through L_1. This variation of grid potential will produce fluctuations in plate current, and it is just possible that these fluctuations in plate current may reinforce the oscillatory current which we have already assumed in the L_1–C_1 circuit.

If this reinforcement of the oscillatory current is great enough, it is possible that the oscillatory current, instead of dying down, may actually build up in amplitude. The form of current is given by

$$i = I_0 \epsilon^{-\alpha t} \sin \omega t \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (70)$$

in which I_0 is the amplitude assumed for the original disturbance (this may be vanishingly small), ω is determined primarily by L_1 and C_1, and α determines how rapidly the current dies away. Because of the reinforcing action mentioned above it is evident that α must involve, among other things, the coupling between grid and plate circuits.

Evidently if α, itself in the above equation is negative, the value of $\epsilon^{-\alpha t}$ increases with time (instead of decreasing) so that the current assumed in L_1–C_1, actually increases. Hence this is the criterion for producing oscillations in the circuit—that the α of eq. (70) must be negative.

A differential equation analysis of the circuit of Fig. 156 shows that for this case α is negative if M, the coefficient of mutual induction between L_1 and L_2, is negative in polarity, and its magnitude satisfies the relation

$$M > \frac{1}{\mu}(L_1 + C_1 R_1 R_p). \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (71)$$

If this equation μ is the amplification factor of the triode, R_1 is the total resistance in the L_1–C_1 circuit, and R_p is the a.c. resistance of the plate circuit of the triode. The biasing battery in the grid circuit prevents this from drawing current as the oscillations start, so that the resistance of the input circuit of the triode does not occur in eq. (71); it is assumed infinite.

Examination of eq. (71) shows that the required mutual induction varies inversely with the amplification factor of the triode, and directly as the factors L_1, C_1, R_1, and R_p.

Let us assume that $\mu = 5$, $L_1 = 100\mu h$, $C_1 = 500\mu\mu f$, $R_1 = 25$ ohms, and $R_p = 4000$ ohms.

The required value of M, in microhenrys, is given by the relation

$$M > \tfrac{1}{5}(100 + 500 \times 10^{-6} \times 25 \times 4000)$$
$$> 30\mu h$$

Hence if $L_1 = L_2$ the coupling must be increased beyond 30 per cent before the circuit will oscillate. If the capacity in the oscillating circuit is increased to $1500\mu\mu f$ and the resistance of the oscillating circuit is increased to 100 ohms the required value of M is calculated to be $140\mu h$, and of course unless L_2 is much greater than L_1 such a value of M is impossible. This means that the circuit could not be made to oscillate.

In carrying out tests to check the relationship given in eq. (71) the resistance R_p may be assumed as one-half of R_{op}, which is given by the quotient of plate voltage by plate current (c.c. values) or it may be obtained by increasing the voltage of the plate by a small amount, ΔE_p, and reading the corresponding increment in plate current ΔI_p. The value of R_p is then given by the expression $\Delta E_p/\Delta I_p$. The resistance R_1 will be made up of the inherent resistances of coil L_1 and condenser C_1, in addition to the high-frequency decade resistance box put in the circuit, for varying R_1. The resistance of L_1 and C_1 in series should be found by the half-current method (Exp. 30).

The values of M for various positions of the coupler L_1–L_2 must be measured, if not known. This is most easily done by bridge measurement, as in Exp. 10. If the calibration of L_1–L_2 connected as a variometer is given, then M can be calculated from the relation that the self-induction of such a variable inductance is given by

$$L_1 + L_2 \pm 2M,$$

and at the mid-position (coils at right angles) the value of self-induction is $L_1 + L_2$.

In carrying out the tests to check eq. (71), two or more different triodes should be used to check the effect of variation in μ, and with one of these triodes, variations of C_1, R_1, and R_p should be carried out. The R_p can most easily be varied by varying the plate voltage.

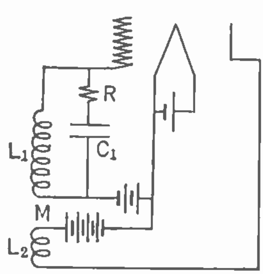


Fig. 157.

It is not advisable to use values of C_1 less than possibly $100\mu\mu f$, because it will be found that oscillations occur for values of M much less than the value calculated from eq. (71). The capacity between the grid and plate of the triode, as well as that between L_1 and L_2, serves as a capacity coupling between plate and grid circuits, thus assisting M, in setting up the oscillatory condition.

It will be found that for the larger values of capacity, experimental and theoretical values of M are in close agreement, but that the experimental value of M is smaller than the theoretical for the smaller values of C_1.

With the circuit set up as in Fig. 157, the value of M required to set up oscillations is given by the relation

$$M > \frac{\mu L_1}{2} - \sqrt{\left(\frac{\mu L_1}{2}\right)^2 - R_1 R_p L_1 C_1}. \quad . \quad . \quad . \quad . \quad (72)$$

Let us suppose $\mu = 5$, $L_1 = 100\mu h$, $R_1 = 25$, $R_p = 4000$, and $C_1 = 1500\mu\mu f$. Then we have for M, in microhenrys

$$M > \frac{500}{2} - \sqrt{\left(\frac{500}{2}\right)^2 - 100 \times 4000 \times 25 \times 1500 \times 10^{-6}}$$

$$> 250 - \sqrt{250^2 - 15{,}000}$$

$$> 32 \mu h$$

Here again, if $L_1 = L_2$ the conditions for oscillation require that the coupling of L_1 and L_2 shall exceed 32 per cent.

In this circuit also it will be found that the agreement of experiment and theory holds good for values of C_1 much greater than the grid-plate capacity of the triode, and the capacity between L_1 and L_2.

A somewhat more complicated circuit is shown in Fig. 158. It is sometimes called the " split coil " circuit. If there is mutual induction between L_1 and L_2 the energy from the plate circuit may reach the grid (for excitation) either through the mutual induction or through the condenser C_1, which, it will be noted, is in parallel with the grid-plate capacity of the triode itself. The condenser C is a large condenser to by-pass the B battery. It is of course in series with C_1 and so must have some effect on the frequency which the circuit generates, but as it will generally be about 1000 times as large as C_1 this latter condenser alone need be considered in frequency calculation.

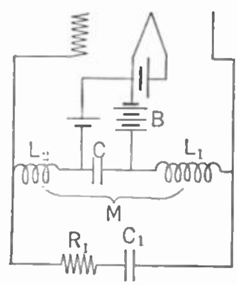


FIG. 158.

According as the relative polarity of L_1 and L_2 is positive or negative the effect of M may be to increase the total effective inductance in the oscillatory circuit or to diminish it. If the oscillatory current flows through L_1 and L_2 in such a direction as to make their magnetic fields additive, the frequency of oscillations, when they occur, is given by the relation

$$\omega = 2\pi f = \frac{1}{\sqrt{(L_1 + L_2 + 2M)C_1}} \quad . \quad . \quad . \quad . \quad (73)$$

and the condition for oscillation is given by

$$\frac{\mu L_2 - L_1 + (\mu - 1)M](L_1 + M)^2}{(L_1 + L_2 + 2M)C_1} > R_1[R_p(L_1 + M) + R_1(L_1 - \mu M)]$$

In case there is no mutual induction, of course these two expressions become much simpler.

$$\omega = \frac{1}{\sqrt{(L_1 + L_2)C_1}} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (74)$$

and

$$\frac{L_1(\mu L_2 - L_1)}{(L_1 + L_2)C_1} > R_1(R_p + R_1). \quad . \quad . \quad . \quad . \quad (75)$$

This last equation shows that C_1 must not be made too large, if oscillations are to be set up, and that the maximum value of C_1 for which oscillations can be sustained decreases as either R_1 or R_p is increased.

In another circuit, called the split condenser method, slightly different methods of plate power supply are generally used. In this circuit, Fig. 159, the oscillatory circuit is made up of one coil and two condensers, in series with each other. This circuit was examined qualitatively in Exp. 27.

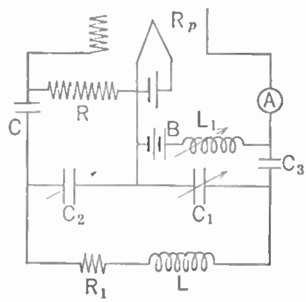


FIG. 159.

In Fig. 159 the coil L_1 is a radio frequency choke coil; it should have a reactance equal to several times the a.c. resistance of the plate circuit of the triode R_p. The condenser C_3 is an insulating condenser to keep the plate voltage (c.c.) from the oscillating circuit; its reactance should be low compared to R_p. Grid condenser C and leak R_g serve to limit the input to the plate circuit; the more violently the L–C_1–C_2 circuit tends to oscillate, the more negative bias does the grid assume.

In testing this circuit, L_1 can be taken at $100\mu h$, the same as in previous tests. Condensers C_1 and C_2 are variable condensers, up to $1500\mu\mu f$. Resistance R_1 is a decade resistance box, wound for radio frequencies.

The student can work out for himself the conditions for oscillation of the circuit, in an approximate fashion, as follows:

For a given plate voltage and plate current a certain power is supplied by the B battery. About 40 per cent of this may be changed to a.c. power, if conditions in the circuit are right. The resistance R_p, across condenser C_1, must be changed to equivalent series resistances and added to R_1.

The power the triode can generate results in a certain current flowing in the oscillating circuit, this current of course depending upon the resistance of the circuit. This current gives a certain drop across C_1 and C_2; that across C_1 must be equal to $1/\sqrt{2}$ of E_b, and that across C_2 must be equal to $2/\mu$ times as much as that across C_1. These are the conditions for maximum power, and must be satisfied if R_1 is large. For low values of R_1 the tube does not require so much power in the oscillating circuit, to satisfy the voltage conditions given above, so that oscillations may occur for quite a wide range of values of C_1 and C_2.

With no added resistance in the oscillatory circuit, and C_1 set at about $300\mu\mu f$, find the range of condenser C_2 which will sustain oscillations. Too small a value of C_2 prevents oscillations because of excessive loss in R_g, and too large a value of C_2 gives no sufficient drop to excite the grid properly.

Find how this range of possible values of C_2 varies as R_1 is increased.

Find how the range varies with E_b and with different values of the choke L_1.

EXPERIMENT 33

Object.—To measure the voltage amplification of a single stage of audio frequency amplification, throughout the audio range. Comparison of resistance, inductance, and transformer repeating circuits. Effect of signal strength on the form of this amplification-frequency curve. Effect of triode plate circuit resistance on form of curve.

Analysis.—One of the main roles of the triode is that of amplifier. The alternating output from the plate circuit is many thousands of times as much as the power required to excite the grid, and the voltage

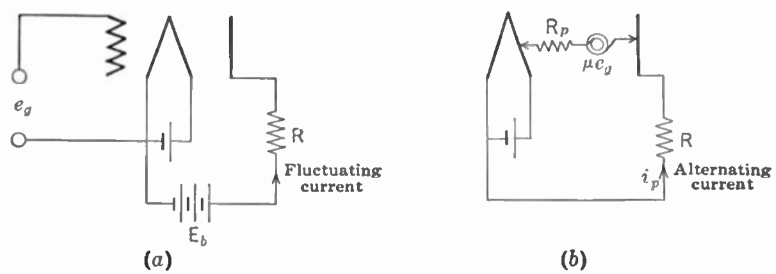


FIG. 160.

form of the output may be made very nearly the same as the voltage impressed on the grid.

The triode (*a*) of Fig. 160 is exactly replaceable by the diode (*b*) of Fig. 160, provided we are not interested in the effect of inter-electrode capacities. So far as its normal function as amplifier is concerned, the diode (*b*) is just the same as the triode (*a*).

In discussing the triode as amplifier we neglect steady voltages and currents, as they have no function directly concerned with amplification.

For so-called resistance amplification the arrangement of circuits is as shown in Fig. 161. The voltage E_g produces in the plate circuit of the first triode a current given by

$$I_p = \mu E_g/(R + R_p)$$

and this current produces a voltage drop across R of $\mu E_g \dfrac{R}{R + R_p}$. By means of the connection scheme shown, most of this voltage is repeated to the grid of the succeeding triode. A leak resistance R of about 1 megohm keeps the grid of the second triode from being "free"; this is very necessary if the amplifier is to act consistently. Of course there is also a resistance inside the triode between the grid and filament, which is in parallel with this grid leak. In determining the value of resistance R_g, these two parallel resistances must be considered.

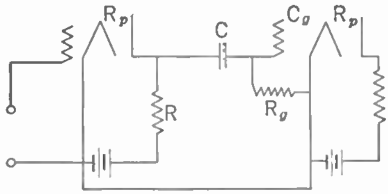


FIG. 161.

The impedance of the C–R_g circuit must be large compared to R, otherwise this circuit will act as a short circuit around R and so reduce the amplification. The fraction of the voltage $\mu E_g \dfrac{R}{R + R_p}$ which appears across the grid of the second triode is given by $\dfrac{R_g}{\sqrt{\left(\dfrac{1}{\omega C}\right)^2 + R_g^2}}$. Unless the frequency is very low, this fraction is sensibly equal to unity, so that the full voltage made available by the first triode is passed along to the second. Of course at very low frequencies $\dfrac{1}{\omega C}$ becomes large compared to R_g, and the voltage amplification of the circuit rapidly falls. At extremely high frequencies (radio frequencies of 10^6 and higher) this scheme fails because the resistance R is short-circuited by the reactance of C and C_g in series. Small as these capacities are, their reactance is low at frequencies of a million and more.

In the inductance repeating scheme the resistance R is replaced by an iron core inductance of many henrys; as much as 50 henrys is frequently employed. In this case the voltage available for passing on to the next triode is $\mu E_g \dfrac{\sqrt{\omega L^2 + R^2}}{\sqrt{\omega L^2 + (R + R_p)^2}}$. In a good coil this is nearly equal to $\mu E_g \dfrac{\omega L}{\sqrt{\omega L^2 + R_p^2}}$, which, for a given magnitude of impedance in the external plate circuit, is higher than when resistance is used. Furthermore, the c.c. resistance of the coil is low, so that practically all of the B battery voltage is available at the plate of the triode. When using resistance for repeating the B battery voltage must be increased in direct proportion to the external resistance. Thus if the R_{op} of the triode is 10,000 ohms and R is equal to 10,000 ohms, the B battery must be of twice the voltage required for inductance repeating, for equal values of R_{op}.

Of course the fraction $\frac{\omega L}{\sqrt{\omega L^2 + R_p^2}}$ falls off with decreasing frequency. If $L = 50$ henrys and R_p is 10,000 ohms the fraction has a value of unity (practically) at 1000 cycles, and only 0.83 at 50 cycles. Thus the lower audio frequencies are not repeated as well as the higher ones.

The most general scheme of audio frequency amplification uses an iron core transformer, with a step-up ratio of about 3 to 1, for repeating, as shown in Fig. 162. The inductance of the primary winding varies from 5 to 50 henrys in various makes, and the secondary inductance is about 10 times as much. It is practically always necessary to use a negative bias on the grid of the triode to which the secondary winding is connected, for reasons to be shown below.

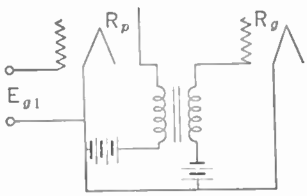


FIG. 162.

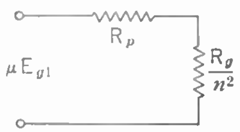


FIG. 163.

By neglecting the primary magnetizing current (temporarily) the circuit of Fig. 162 can be simplified to that of Fig. 163, in which it has been assumed that the transformer ratio is n. The voltage across the hypothetical resistance R_g/n^2 is equal to $E_g \frac{\mu R_g}{n^2 R_p + R_g}$, and the voltage across the input circuit of the second triode is n times this much. The *voltage amplification per stage* is therefore

$$\frac{E_{g_1}}{E_{g_2}} = \frac{\mu n R_g}{n^2 R_p + R_g} = \frac{\mu n a}{n^2 + a} \quad \ldots \ldots \ldots \quad (76)$$

in which $a = R_g/R_p$.

Analysis of equation (76) shows that if we assume R_g, R_p, and μ fixed the voltage ratio $\frac{E_{g_1}}{E_{g_2}}$ is a maximum for a transformer ratio

$n = \sqrt{R_g/R_p}$. This ideal ratio is never used because the magnetizing current of the primary is not of negligible value. A diagram which represents the conditions of Fig. 162 better than the simple Fig. 163 is given in Fig. 164, in which X is the no load reactance of the primary winding. Evidently X should be high compared to R_g/n^2, if the above theory is to be at all applicable. This means that the primary inductance must be high, and of course the secondary inductance is n^2 times as much. With a very high secondary inductance another trouble is encountered, owing to the fact that the winding itself has a not negligible internal capacity, and the triode, wiring, etc., connected to its secondary terminals also have considerable capacity.

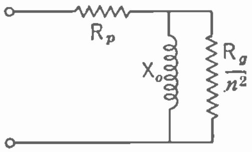


FIG. 164.

The total capacity connected to the secondary winding together with the mutual *leakage reactance* of the transformer determine a resonant frequency, and above this resonant frequency the capacity acts to short-circuit the secondary terminals of the transformer, thus cutting down the ratio E_{g_2}/E_{g_1} to a very small value.

The mutual leakage inductance of the transformer is given by the relation $L_m = (1 - k)\sqrt{L_1L_2}$, in which k is the coefficient of coupling of the primary and secondary windings. This factor is seldom as much as 0.9, with transformers as ordinarily wound, and generally much less. Suppose $L_1 = 50$ henrys, $n = 5$, then $L_2 = 1250$ henrys and $L_m = 0.1\sqrt{50 \times 1250} = 25$ henrys.

If now the total effective capacity across the secondary winding is $50\mu\mu f$, the resonant frequency due to this leakage reactance is given by

$$f = \frac{10^6}{2\pi\sqrt{25 \times 50}} = 4600 \text{ cycles.}$$

This transformer would then give a resonant rise in amplification at this frequency, and above this value of frequency the amplification would rapidly fall to zero.

At the lower limit of the audio range the action of X_o (Fig. 164) acts to cut down the voltage across the resistance R_g/n^2 and so cuts down the voltage E_{g2}. Thus at the lower and upper limits of the audio frequency range the amount of amplification obtainable is seriously limited. By coupling the primary and secondary coils tightly, taking care to have as little capacity in the secondary circuit as possible, using a high primary inductance and low ratio of turns, these defects in the amplification curve are somewhat remedied.

In Fig. 165 curve B shows the amplification obtained from a poorly designed transformer, and curve A that from a lower ratio, better-designed transformer. In present practice it is not attempted to amplify audio frequencies higher than about 7000 cycles.

The resistance of the input circuit of a triode decreases very rapidly with increase in the amplitude of the impressed voltage. Thus, if the voltage impressed on the input circuit of the first triode E_{g1} is held fixed as the frequency is varied through the audio range, it happens that the input resistance R_g of the second triode is much less for some frequencies than for others. Now the frequencies which amplify the most are evidently those at which R_g assumes its lower values, but we know that if R_g decreases the voltage amplification also decreases. (See eq. 76.) Hence we may expect the form of the amplification curve to be different at different signal strengths, and such proves to be the fact. In

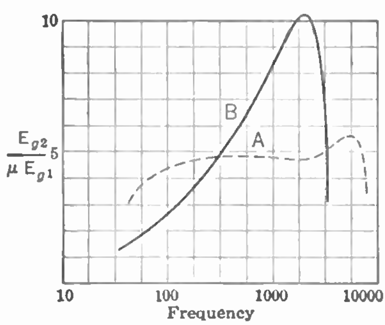


FIG. 165.

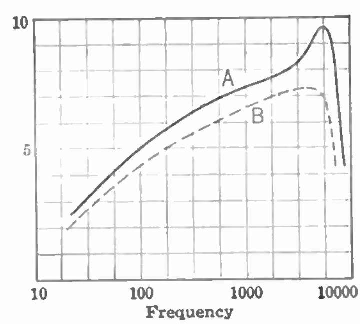


FIG. 166.

Fig. 166 are shown two experimental curves on the same transformer; curve A is for a weak signal and curve B was for a strong signal.

Because of this dependence of amplification upon the signal strength it is necessary that tests be carried out with the same signal strength if the results are to be comparable. It has been recommended to use 0.075 volt as the standard signal for tests of this kind.

By referring to eq. (76) it is evident that the amount of voltage amplification obtained, as well as its variation with frequency, will depend upon R_p, the plate circuit resistance in series with the primary winding of the transformer. As the plate circuit resistance of a triode depends upon the plate voltage used, it is evident that more amplification, and more uniform amplification, will be obtained if the voltage of the B battery is increased.

However, there may be a very definite limit to this improvement in performance. More plate voltage means more plate current, and this means more c.c. magnetization of the iron core of the transformer. But the a.c. permeability of the transformer core (and hence the reactance of the primary winding) drops rapidly as the core is magnetized more strongly by the continuous current, and it may be that in a transformer having many turns in the primary winding this effect more than offsets the advantage which should be gained by increasing the plate voltage.

This decrease in a.c. permeability as the continuous current magnetic motive force is increased will be quantitatively investigated in Exp. 36.

It will be appreciated at once that it is impossible to measure the voltage amplification called for in this experiment by using ordinary voltmeters; the highest-resistance a.c. voltmeter available has a resistance of about 2000 ohms per volt, and quite evidently connecting such a voltmeter across the secondary of the transformer being tested would practically short-circuit it! It might seem that a low reading electrostatic voltmeter would serve, but even this is impossible. The amount of capacity in such a meter together with the capacity of its connecting wires would completely alter the amplification from the value existing when the voltmeter is not connected.

The only feasible voltmeter to use for making the voltage measurements called for in this test is the triode voltmeter; its resistance is several hundred thousand ohms, and its internal capacity is very low. Moreover, it so happens that the voltages which are to be measured in this test are just those for which the ordinary triode voltmeter is suitable, namely, up to about 3 volts.

From what has been said regarding the change in amplification with changes in R_g, etc., it appears that even connecting a 500,000-ohm voltmeter across the secondary terminals may seriously affect the performance of the amplifier. However, in normal use, a stage of amplification always does have a triode connected to the secondary terminals of the transformer, so we will use this triode itself as a voltmeter, thus measuring the performance of the amplifier stage when it is connected exactly as when being used.

Although there is on the market an amplification measuring set, it will pay the student to design and set up for himself a proper circuit and try to get consistent amplification values. The input voltage E_g is obtained by a potentiometer connection to the output circuit of an audio frequency oscillator, this having a frequency adjustment as wide as the range over which the amplification is to be measured. A thermocouple ammeter measures the current through the resistance, and this, multiplied by the resistance across which the signal voltage is taken, gives the magnitude of the signal. As mentioned before, this should be held constant at 0.075 volt, or other convenient low value.

With no grid condenser in the input circuit of this second triode, the plate current will increase with the voltage impressed on its grid filament circuit. This triode, instead of being calibrated, as a voltmeter, is used simply as a comparator. Its grid is transferred from the transformer secondary to the input signal potentiometer, and this is adjusted to give the same plate current as when the grid was connected to the transformer secondary. This new value of resistance, compared to the value from which the input signal was obtained, gives the voltage amplification per stage (triode + transformer). The performance of the transformer is obtained by dividing this value by μ, the triode amplification factor.

It will be found very difficult to get consistent results from this

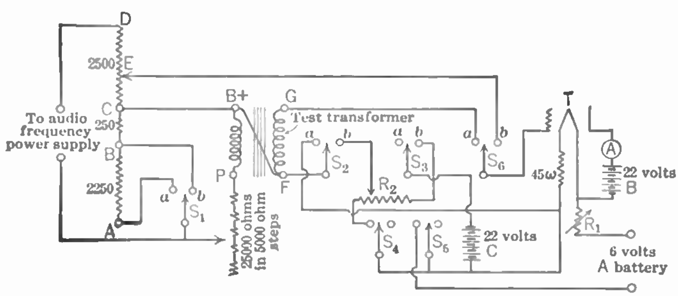


FIG. 167.

test until much experience has been gained. A slight rearrangement of the connecting wires, or variation in their length, may change the results as much as 50 per cent at the higher frequencies. The results must not be regarded as reliable until they can be repeated with a new set-up of the apparatus, in which the various parts and connections have been rearranged. As long as a slight rearrangement of wires changes the results they are undoubtedly unreliable.

The amplification of several transformers, under various conditions, is then to be measured with a commercial outfit (that made by the General Radio Co.). It is wired as in Fig. 167; by its use several transformers can be tested in one afternoon.

The transformer to be tested is connected as shown at *B G P F*, these symbols being the markings put by the manufacturer on the transformer terminals. When connected in an amplifying set the terminals *B* and $-F$ are at ground potential (in so far as a.c. is concerned), so they are connected together in this apparatus.

The triode T is of the 99 type, having $22\frac{1}{2}$ volts, plate battery B, and microammeter for measuring its plate current. A 45-ohm resistance in the negative leg of the A battery gives sufficient bias to prevent the grid from drawing current, unless the impressed voltage is several volts. When this occurs, the extra battery C, through its potentiometer R_2, affords the extra grid bias required.

Instead of putting the primary winding of the transformer in an actual plate circuit, the resistance of the plate circuit is represented by the resistance R, which is adjustable, in steps of 5000 ohms, up to 25,000 ohms. Amplifier tubes (of the ordinary type) have values of R_p from 5000 to 10,000 ohms, whereas the detector may have an R_p up to 25,000 ohms.

The audio frequency oscillator (of adjustable frequency and voltage) is connected across the resistance shown at the left of the figure to points D–B or D–A, according as switch S_1 is thrown to b or a. When testing transformers it is always thrown to point b; then it is apparent that the drop across resistance C–B is impressed on the transformer and resistance R in series.

The scale of triode T is not calibrated as a voltmeter but one point is calibrated for one volt.

The resistance D–C is calibrated, really with the 250-ohm value as unity. With S_6 thrown to b, and slider E set to give 250 ohms between C and E, the voltage impressed on the triode is the same as that impressed on the transformer primary circuit. Hence if the voltage from the A.F. power supply is adjusted until the triode reads "one volt" (the zero adjustment of the ammeter having been set by varying resistance R_1), then there is one volt signal impressed on the transformer primary circuit. If the slider E is set to include all of resistance CD, and the power supply is adjusted to make the triode read "one volt," then there is evidently 0.1 volt impressed on the transformer primary circuit. This then is the signal strength adjustment for the set.

Switch S_6 is then thrown to position A, and the reading of the triode voltmeter noted. If it goes off scale it is brought back by use of additional bias, obtained from battery C and resistance R_2 through switches S_2 and S_3. Switch S_6 is now thrown to b and contact E is moved until the ammeter A reads the same as when the triode was connected to the secondary of the transformer (switch S_6 on a).

When this equality of reading has been obtained, the resistance CE, in terms of 250 ohms, gives the amplification of the transformer.

In case a resistance repeater unit or inductance repeater unit is to be tested, the amplification factor will always be less than 1. In this case the drop across the resistance C–A is impressed on the unit by throwing switch S_1 to a. The amplification read from the position of slider E, after balance has been obtained, is then multiplied by 0.1.

With normal conditions as regards signal strength, scheme of connections, plate resistance, etc., get the amplification curve of three or four audio frequency transformers, preferably two or three of them built a few years ago, and one good one of modern design.

With one or two of the transformers, try the effect of connecting the secondary winding backward, that is, reverse the connections of terminals G and $-F$.

With one of the transformers get amplification curves, when the transformer secondary is shunted by about three resistances of values representing possible resistances of the input circuit of a triode, say, 10^6, 3×10^5, and 10^5 ohms. Try the effect of a $50 \mu\mu f$ condenser added across the secondary terminals.

With one or more of the transformers, try the effect of c.c. magnetization on the transformer characteristics, by connecting a suitable battery, of a few volts, in series with the primary coil of the transformer. Read the current in the primary winding with a milliammeter, and get the amplification curve for two or more values of plate current, such as might flow in the primary coil when in an actual amplifier. (Battery must be removed when checking the input voltage.)

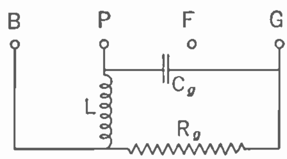


Fig. 168.

If time permits, get one curve for resistance amplification, and another for an inductance repeating unit. For these units connection is made to the test apparatus as shown in Fig. 168.

Plot all curves on semi-logarithmic cross-section paper, plotting frequency on the logarithmic scale. The use of this scale is practically universal in representing the characteristics of sound apparatus.

EXPERIMENT 34

Object.—Calibration of a wave meter. Effect of resonance indicating device on calibration. Use of standard wave meter, and use of standard L and C for calibration. Checking one or more points by piezo electric crystal, and by harmonics. Check standard wave meter against Bureau of Standards " standard signals."

Analysis.—A wave meter, or rather frequency meter, is perhaps the most useful meter in the radio laboratory; it is used in making all kinds of radio frequency measurements.

Primarily it is nothing but a fixed coil of low R.F. resistance connected to a well-constructed, low loss, variable condenser. Generally some sort of current-indicating device is incorporated in the meter, to show when its current is a maximum. For many purposes a current indicator only is necessary; for others a calibrated ammeter is required.

A standard wave meter is always of very rugged construction, so that its calibration may be permanent. Its resonance indicator is a small thermocouple ammeter rigidly mounted inside the shielded box containing the condenser, if a resonance indicator is used. Frequently no resonance indicator of any kind is used, the meter being tuned *by its effect on the measured circuit.* A laboratory wave meter sometimes has phones and crystal as one method of detecting resonance, but quite evidently having the loose flexible wires of a phone connected to the wave meter circuit does not make for accurate or permanent setting. Furthermore, it is evident that the calibration will be different with phones and crystal across the tuning condenser than when they are not used. This error will be greater as the capacity of the wave meter condenser is set for its lower values.

Variable condensers are available which are calibrated to a small fraction of 1 per cent, and fixed standard inductances are available the values of which are given with the same precision. Thus with the use of these two laboratory standards (which of course can be checked against others) a frequency meter can be made which has a precision of a small fraction of 1 per cent. Of course, short stiff wires should be used in connecting the coil to the condenser. As no resonance indicator is used, the wave meter must be tuned by noting the effect on the circuit being measured.

Precision wave meters are available which are guaranteed to be accurate to within 0.25 per cent.

Piezo electric crystal controlled triode oscillators will generate a frequency which is remarkably constant. A simple circuit utilizing this principle is shown in Fig. 169. The small crystal disc is held loosely between two plates as shown at *P*. When the *L–C* circuit is tuned approximately to the natural frequency of mechanical vibration of the quartz disc, the circuit oscillates at exactly the frequency of the disc. The frequency of the disc is generally given by the manufacturer to 0.1 per cent or better.

The operator knows when the circuit is in oscillation by the reading of meter *A*; the plate current falls suddenly when oscillations start. Such an oscillator is used to set the frequency of a laboratory oscillator, by listening in telephone receiver *T*. The beat note is heard, until the laboratory oscillator has exactly the same frequency as the piezo electric oscillator. The wave meter to be calibrated is then excited from the laboratory oscillator circuit.

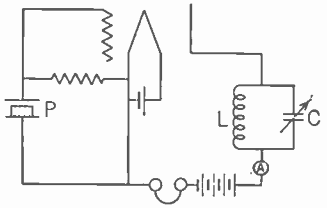


FIG. 169.

The piezo electric oscillator generates a whole series of harmonics; if the plate current of the oscillator is made to induce voltage on the grid of another oscillator, the frequency of which can be made two, three, or higher multiples of the crystal frequency, beats can be heard between the harmonic frequency of the crystal and the frequency of this oscillator. Thus the one calibrated frequency of the crystal may be made to yield a dozen or more other standard frequencies, in the higher frequency ranges.

The Bureau of Standards sends out a program of standard frequencies from their station WWV at regular periods. This broadcast of standard frequencies is really the best way to calibrate a wave meter, or at least check a few points on the calibration curve. This is best done by listening for the continuous wave signals sent out, with an oscillating receiver; if the frequency of this receiver is set to give " zero beat " with the signals from WWV it is oscillating at exactly the same frequency as WWV, which frequency is certified by the Bureau of Standards to four places. This oscillating receiver is then used to check the wave meter to be calibrated; when the wave meter is tuned to the receiver, and coupled sufficiently close, it will make the receiver stop oscillating when it is tuned to the same frequency as the receiver, as shown in Exp. 31.

Calibrate one or more of the laboratory wave meters by comparison with the laboratory standard. Unless there is some mechanical defect in the variable condenser of the wave meter being tested, its capacity will vary smoothly with its setting. The calibration of the wave meter must therefore be a smooth curve. The various calibration points obtained are therefore to be connected by a smooth curve, and the check points obtained by laboratory standard inductance and condenser, by piezo crystal and its harmonics, and Bureau of Standard signals should lie on this smooth calibration curve.

EXPERIMENT 35

Object.—To determine the decrement of a wave meter, by variations of its capacity, and by variations of impressed frequency.

Analysis.—The sharpness of resonance of a wave meter is determined entirely by the ratio of its resistance to the reactance of either coil or condenser; the less this ratio, the sharper is the tuning of the meter.

In determining the decrement of the meter it is necessary to read the current in the wave meter circuit under different conditions, hence it is not possible to determine directly the decrement of a meter which has no current-indicating device in its circuit.

In determining the decrement of a damped oscillation (a necessary task in connection with every spark transmitter) the reading of the wave meter ammeter is taken for various settings of the condenser. The resonance curve so obtained has its shape determined by the sum of the decrement of the damped oscillation and that of the wave meter itself; the decrement of the oscillations is found by subtracting from this combination decrement the decrement of the wave meter, which is presumably known. In Fig. 170 are shown two resonance curves, one obtained by varying the frequency impressed on the wave meter, with the wave meter setting and impressed voltage held constant. In the other the impressed voltage and frequency are held constant, and the setting of the wave meter condenser is varied. The square of the current in the wave meter circuit is used as ordinate for both curves; as the meter is generally a thermocouple—c.c. galvanometer combination, the readings of the galvanometer plotted as ordinate do represent the (current)^2, as proved in Exp. 14.

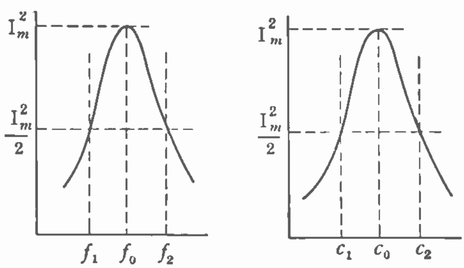


FIG. 170.

In the first curve three frequencies are noted: f_0 which gives the maximum current I_m^2, and two others (one on each side of f_0) which give one-half of this (maximum current)^2. In other words, if the galvanometer of the thermocouple ammeter reads 50 at resonance, the frequencies f_1 and f_2 are those which make the galvanometer read 25.

The alternating current generated by a triode or other high-frequency generator has no decrement; its successive cycles all have the same amplitude. Hence if the resonance curve of Fig. 170 is obtained from a wave meter excited by a triode oscillator there is only one decrement affecting the shape of the curve, namely that of the wave meter itself.

The decrement, generally indicated by the letter δ, is given by the relation

$$\delta = \pi \frac{f_2 - f_1}{f_0}. \qquad (77)$$

For a good wave meter this should not exceed a few hundredths.

If now the frequency and voltage have been held constant, and the resonance curve obtained by varying the wave meter condenser (second curve of Fig. 170), the decrement is given by

$$\delta = \frac{\pi}{2} \frac{C_2 - C_1}{C_0}. \qquad (78)$$

In using the relation of eq. (77) the wave meter is coupled to an oscillating triode and tuned to it. The coupling is then increased until the wave meter ammeter reads well up on its scale. The frequency of the oscillator is then varied in small steps, and the current in the wave meter under test is read. For each step the frequency of the oscillator is read (by another wave meter or from the calibration curve of the oscillator, if this is available) and the current of the oscillator noted. For the very small range of frequencies required to plot the required resonance curve, the current of the oscillator will remain essentially constant under ordinary conditions.

Of course even if the current does remain constant, the voltage induced in the wave meter under test will not be quite constant, as it should be in getting a resonance curve. The induced voltage is equal to $2\pi fMI$, so that really I should go down as f goes up, to keep the induced voltage constant. However, as the whole range of frequency required is only 1 or 2 per cent, the error introduced is quite small.

In getting the resonance curve, in which current and wave meter capacity are the coordinates, the procedure is somewhat different. The calibration curve of the wave meter condenser is first measured, if it is not available.

With the wave meter set for that frequency for which the decrement is desired, the triode oscillator is coupled to the wave meter and adjusted to resonance with the wave meter. The coupling is then adjusted to give a reading on the wave meter ammeter well up on the scale. With the frequency and current of the oscillator constant, and coupling at the value fixed as above, the wave meter condenser is varied in small steps, and the wave meter ammeter read. Sufficient points are obtained to get a well-defined resonance curve.

In addition to the definitions of decrement given in eqs. (77) and (78), it may also be defined by the relation $\delta = R/2fL$, in which R is the resistance of the wave meter circuit, f the frequency at which the decrement is desired, and L is the inductance of the wave meter coil.

This expression for δ leads one to believe that the decrement varies inversely as the frequency impressed on the wave meter, but this is by no means the fact. The high-frequency resistance of a coil increases with the frequency, not linearly to be sure, but still it does increase with frequency, and sometimes more rapidly than the frequency. As wave meter coils are generally built, the resistance does increase nearly as the frequency, so that the decrement, when using a given coil, is nearly independent of the frequency.

By the variations of wave meter condenser get the decrement of one of the laboratory wave meters for about six points, distributed throughout the range of the condenser. Do this for two or three different coils, covering different wave length ranges.

For one or two of the points check the value of decrement obtained by this method with that obtained by varying the impressed frequency.

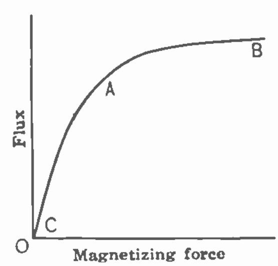
EXPERIMENT 36

Object.—To measure the a.c. permeability of iron at audio frequencies. Variations of permeability with frequency, with strength of a.c. magnetomotive force, and for a fixed a.c. magnetomotive force with various c.c. magnetomotive forces superimposed. Effect of air gap.

Analysis.—Iron core coils are extensively used in radio apparatus, as witness the transformer or choke coil of the audio frequency amplifier, the modulation coils of a radio telephone transmitter, the choke coils used in the filters of power supply units, etc.

All these coils have iron core coils, not because a higher ratio of reactance to resistance can be obtained than is possible with air core coils, but because a high reactance can be obtained with smaller weight and volume of coil.

Unless a special study has been made of the action of iron for weak alternating magnetomotive forces, the student is likely to have very erroneous ideas regarding its behavior. The claims of some of the manufacturers of electric iron, for permeability, etc., are exaggerated as much as five times; the values of permeability they claim in their catalogs are five times as much as the actual values. This over-statement may result in very poor performance of transformers, choke coils, etc., if the radio designer depends upon the catalog values of permeability.



(*a*)

FIG. 171.

In Fig. 171 is shown the *B–H* curve of ordinary electric iron; if the ratio of *B* to *H* is taken for some such point as *A*, the permeability will be found about 3000, and values as high as this are sometimes claimed for the a.c. permeability. This value of 3000 is a proper one for the designer of c.c. machinery, but the radio designer must be content with a value about one-fifth of this.

The a.c. permeability must be defined by the slope of the *B–H* loop, when the iron is carried through a magnetic cycle, as indicated in Fig. 172. Now the slope of this *B–H* loop varies with the intensity of the magnetization, about as shown in Fig. 173. For weak m.m.f.'s the slope (and hence μ) is small, as for loop *A*. For stronger m.m.f.'s (as loop *B*) the slope is higher, and for still greater m.m.f.'s (loop *C*) it is smaller again. This phenomenon results in an a.c. permeability variation about as shown in Fig. 174, in which the maximum value is about 1500 or 2000. In most of the uses to which the radio designer puts iron he must be content with the low values, for the weak m.m.f.'s.

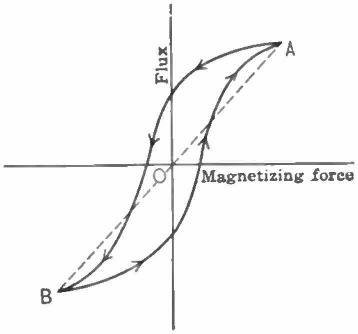


FIG. 172.

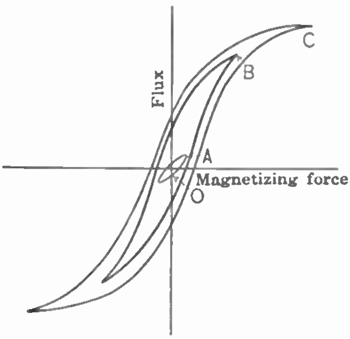


FIG. 173.

The value of permeability is further complicated by the continuous m.m.f. to which the iron is subjected, while the alternating m.m.f. is active. Thus the current in the primary winding of an audio frequency transformer is about as shown in Fig. 175; the actual current is a unidirectional fluctuating one, which can be divided into its two

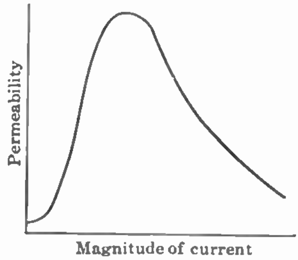


FIG. 174.

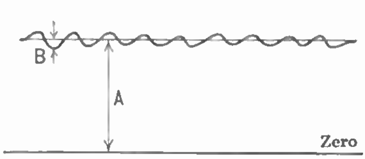


FIG. 175.

components. The c.c. component has the value *A*, and the alternating component has the amplitude *B*.

The a.c. permeability is much lower when the continuous m.m.f. is acting than when it is absent, and it becomes continually lower as the strength of the c.c. m.m.f. is increased. This is brought out in the experimental results given in Fig. 176 which shows typical behavior of the iron used in audio transformers a few years back. Transformers as built today use a more suitable grade of iron, in which higher permeability is available, and for which the c.c. m.m.f. has a less deleterious effect on the a.c. permeability.

The amount of c.c. m.m.f. acting in the cores of A.F. transformers is not sufficiently high to affect the a.c. permeability to any great extent, but the cores of choke coils present a different problem. Here the c.c. m.m.f. is so great that the a.c. permeability may be reduced to an extremely low value. In this case the a.c. permeability is actually increased by putting an air gap in the iron core! This air gap cuts down the continuous flux, for a given c.c. m.m.f., thus making the iron

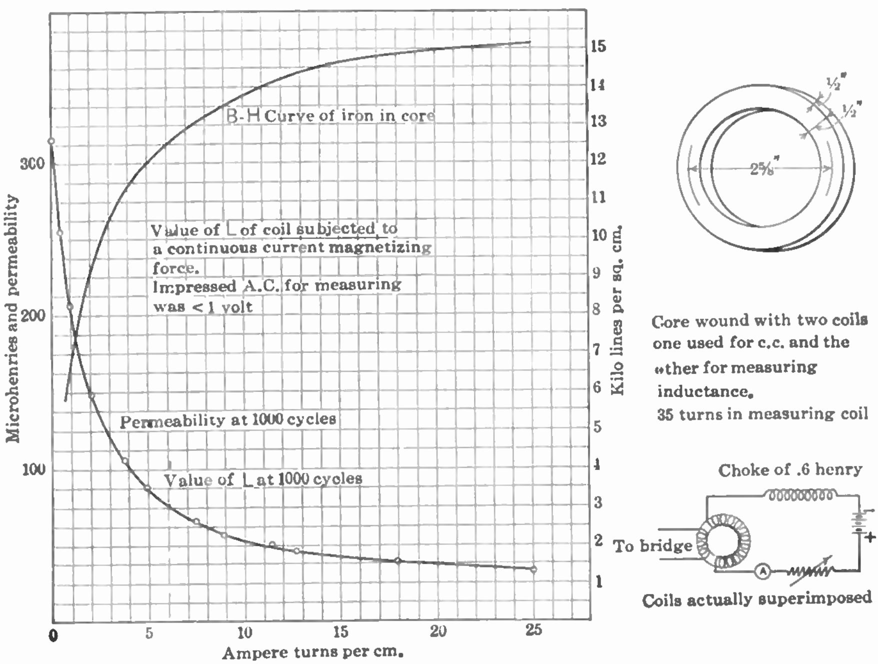


Fig. 176.

operate lower down on the B–H curve and resulting in a high a.c. permeability. This effect is shown in Fig. 177, in which the a.c. inductance of an iron core coil is given, for various air gaps and continuous m.m.f.'s.

It sometimes happens that an alternating current is passed through an iron core coil, which is subjected to another alternating current m.m.f., of another frequency, at the same time. Thus suppose that measurements of inductance and effective resistance of a coil are made at 20 kc., when the iron is subjected to a 60-cycle m.m.f. This measurement is difficult to make because the 20 kc. R and L vary continuously, having greatly different values at different parts of the cycle of the 60-cycle current. In Fig. 178 are shown typical results from a measurement of this kind. A commutator, run synchronously from the 60-cycle supply connected the Wheatstone bridge (by which the L

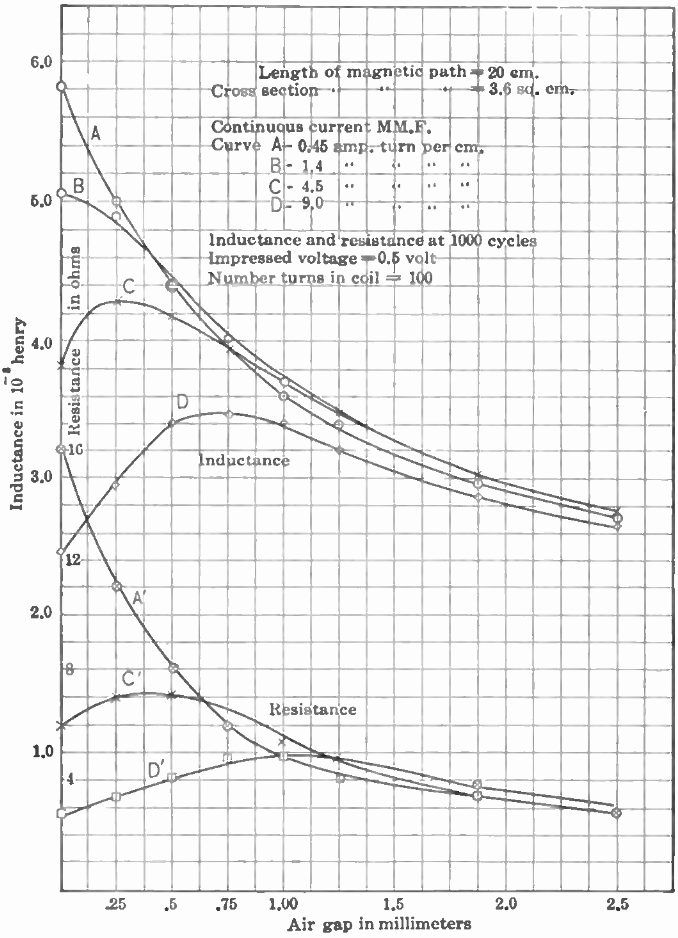


Fig. 177.

and R measurements were made) to the 20-kc. coil at successively different parts of the 60-cycle current, and the bridge balanced.

It is thus evident that this phenomenon offers the possibility of modulating the amplitude of the 20-kc. current, by the 60-cycle current. Such a scheme has been used to modulate the high-frequency current in an antenna, by the voice current in a radio telephone channel.

It can be expected, of course, that the self-induction of an iron core coil will decrease with an increase in frequency, because of the back magnetomotive force set up by the eddy currents in the iron core. With increase in frequency, these currents increase in amplitude, and lag more in phase, both of these actions giving more back m.m.f. to the eddy currents. Thus the flux set up by the coil being measured is increasingly less, per ampere of current in the exciting coil, than it would be if there were no eddy currents, and this means a smaller measured value of self-induction.

It is convenient to have a piece of apparatus, on which the laminations to be tested can be conveniently clamped down, fitted with two

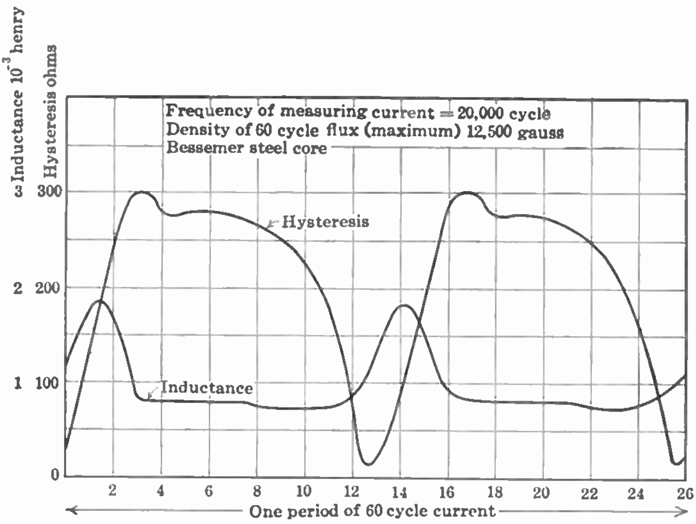


FIG. 178.

coils, through which the laminations are passed. One of these coils is used for carrying the continuous current which gives the required c.c. m.m.f., and the other coil serves for making the a.c. measurements. A coil having sufficient turns to give about 50 millihenrys of inductance is well suited for ordinary measuring apparatus.

The c.c. coil, when used, is connected to a battery through a variable resistance, ammeter, and air core choke coil in series. The reactance of this choke coil should be so high that the c.c. circuit has no appreciable effect on the measured value of self-induction of the a.c. coil. This can be tested by determining the L of the a.c. coil with the c.c. circuit open and then redetermining it with the c.c. circuit closed with no battery in the circuit. There should be no difference in the two values so determined.

If the effect of air gap is to be studied, suitable provisions must be made to clamp the laminations tightly together, with the exception of one corner having a butt joint. This should have smooth faces, fitting snugly against each other, with means for opening the gap to any desired length.

The laminations which are used for the other test are not well suited for this purpose, because all the corners of these laminations should be arranged in lap joints, otherwise the value of permeability determined will be lower than it should be. A butt joint always introduces a considerable reluctance in the magnetic circuit, no matter how well it is made.

The most convenient method of measuring the inductance and effective resistance of the coil is by the alternating current bridge, as it was used in Exp. 10. In this test, as in Exp. 10, the resistance of the variable standard inductance must be known for various positions and frequencies.

The "three voltmeter method" indicated in Fig. 179 is also a feasible scheme for measuring the resistance and inductance of the coil

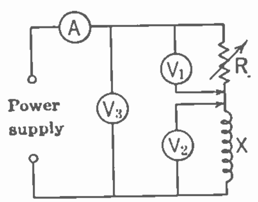


FIG. 179.

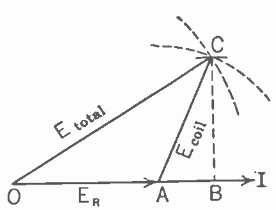


FIG. 180.

under test. An a.c. voltmeter, of sufficiently high resistance, is used to measure the drop across the resistance box R, coil X, and across the two in series. The two readings across R and X must add vectorially to give that across the two in series, as shown in Fig. 180. The decade resistance box R should be varied until it gives a drop about the same as that across the coil. The drop across R is in phase with the current in the circuit, and is plotted to some convenient scale at OA in Fig. 180. Then with the coil voltage as radius and A as center an arc is struck, and with the voltage V_3 (Fig. 179) as radius and O as center another arc is struck. The point where the two arcs intersect, at C, forms the third corner of the triangle made up of voltages V_1, V_2, and V_3.

The component $C\,B$ is equal to the $2\pi fLI$ of the coil, and the component $A\,B$ is the $R\,I$ of the coil. Knowing I from the ammeter A, the X and R of the coil can be calculated.

The suitability of the voltmeter used for making this test is shown by noticing the reading of A, with the voltmeter connected across the coil and then without it. If the two currents differ appreciably the voltmeter has too low a resistance to be used.

Another method utilizes the idea that a series condenser may neutralize the reactance of the coil to any degree desired. The simple scheme of Fig. 181 enables the L of the coil to be determined if only C has a wide enough range of values, and is of low power factor.

With the switch S thrown to a the current is read. Then with the switch thrown to b, C is varied until the current is the same as it was before. This adjustment is checked by throwing the switch back and forth and changing the condenser C until no perceptible change in current occurs when the switch is thrown.

The reactance of the condenser C is then just *twice* the reactance of the coil L; with the switch thrown to a there is an inductive reactance present, and with the switch thrown to b there is an equal reactance. However, a little thought will show that the only possible reactance which C may have, to give the circuit the same impedance it had with the coil alone, is just equal to twice the coil reactance.

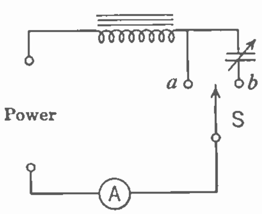


Fig. 181.

Knowing the capacity and frequency, the reactance of the coil is readily calculated, and from this the inductance is found. If this scheme is to give the resistance of the coil as well, it is necessary to measure the voltage impressed on the circuit. Knowing this, and the current, the impedance can be calculated, and from this, knowing the reactance, the resistance is obtained.

Using the bridge method, determine the permeability and effective resistance of the sample of transformer laminations, as a function of a.c. magnetomotive force, the frequency being constant at some suitable value, and the c.c. circuit being open.

Then find how permeability and effective resistance vary with frequency throughout the audible range, by keeping the alternating current, used for measuring, constant in amplitude, as its frequency is varied through as wide a range as possible. In this test, too, the c.c. circuit is to be open.

Then with constant amplitude and frequency of a.c. used in making the measurement, determine the inductance and resistance for gradually increasing c.c. magnetization, up to values sufficient to saturate the iron. In this test care should be observed to increase the c.c. m.m.f. *just up to the point* to be measured; if this value is exceeded, and the current is decreased to the desired value, the value of L determined will be different than it should be. It will be remembered that the same precaution must be taken in getting the magnetization curve of a generator, and for the same reason.

For some values of frequency and current (no c.c.) covered by the foregoing tests, try to check the values by the three voltmeter method, and by the " constant impedance " method.

With fixed value of alternating current, measure the self-induction of the iron core coil as the c.c. m.m.f. is varied, with various air gaps in the magnetic circuit. Keep the frequency constant, at some suitable value.

EXPERIMENT 37

Object.—Study of the effect of dead ends on the high-frequency resistance and inductance of coils.

Analysis.—The radio experimenter has continual need of coils of various inductances, so it is quite a natural tendency to make coils of considerably more inductance than the smallest desired value, and put taps on this larger coil at suitable points. In general, this practice is bad, as it very considerably increases the resistances of the smaller inductances over what they should be, and makes their inductance vary rapidly with frequency, especially at the higher values.

In Fig. 182 we have represented a single-layer solenoid of, say, 100 turns, having about 1 millihenry of inductance. At C there is a tap, giving, say, 25 turns between A and C, and possibly 200 microhenrys of inductance.

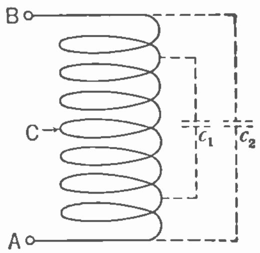


Fig. 182.

Every coil has internal capacity, one part acting as one plate of a condenser and another part (at the other end of the coil) acting like the other plate. This is illustrated in Fig. 182 by the dotted line. Whereas only two of these condensers are indicated it will be understood that the actual internal capacity is complex in nature, there being capacity between adjacent turns, between one turn and every other, etc. There is capacity through the insulation between wires, capacity through the material of the spool on which the coil is wound, and capacity in the air both inside and outside the solenoid.

Now when the coil $A\ C$ is used the whole coil $A\ B$ is excited, and the voltage across $A\ B$ is much greater than the voltage being impressed on the coil terminals A–C. This high voltage of course sends a correspondingly large charging current into the condenser C_2, and the current must circulate through the coil winding and thus produce loss. Furthermore there is always considerable dielectric loss in the internal capacity C_2.

Besides this extra loss in the coil, this unused part has an appreciable effect on the inductance of the coil $A\ C$. The current through the unused part $C\ B$, being a leading current, and this part being coupled to the used part $A\ C$, the effective inductance of the part $A\ C$ may be very much increased; this effect was studied quantitatively in Exp. 5. Thus the effect of the unused portion is to increase the inductance of the tapped portion $A\ C$, and to increase its resistance.

These effects both become very large at high frequencies. Thus let us suppose the internal capacity of the coil A–B is $5 \mu\mu f$, this being a reasonable value for a 4-inch diameter solenoid. This much capacity, acting with the inductance of the coil $A\ B$, will give resonance for a frequency of 1600 kc. Now if it is attempted to use the 200 microhenry tap at this frequency it will be found that instead of having a few ohms, as it should have, it has perhaps 100 or 200 ohms, and so is absolutely useless. Furthermore, for frequencies just below this resonance value, instead of having 200 microhenrys the coil will have a very large inductance, which varies greatly with small changes in frequency. These effects will be appreciated by again studying the results of Exp. 5.

Still another effect will be soon discovered when working with such a tapped coil at high frequencies. When the 200-microhenry tap is tuned with a condenser, it will be found that the circuit responds to two frequencies; as though it had two resonant frequencies; of course it really does have two resonant frequencies, as was shown in Exp. 9.

Using the method of Exp. 29 measure the resistance and inductance of a 25-turn solenoid throughout its useful frequency range. Next measure the resistance of a 25-turn tap on a 50-turn solenoid, this being wound of the same size wire and on the same diameter spool, of the same material, as used in winding the 25-turn solenoid.

Next measure the resistance and inductance of a 25-turn tap on a 100-turn solenoid, this being made in similar fashion to the 50-turn solenoid.

It will be found that at low frequencies the 25-turn coil and the 25-turn tap on the larger coils all have the same resistance and inductance, but that with increasing frequency the tapped portions of the larger coils have increasingly greater resistance and inductance than the 25-turn coil.

At some frequency where the 25-turn tap on the 100-turn coil shows an excessive resistance, short-circuit the unused part of the coil, that is, the 75 turns. It may found that the resistance of the 25-turn tap has actually *decreased*, with the short circuit present. This is due to the fact that high voltage cannot build up across the unused portion with the short circuit present, and this means that the current flowing in the 75-turn portion of the coils is actually less when short-circuited than when open.

EXPERIMENT 38

Object.—Measurement of shielding by different substances and thickness of material.

Analysis.—With the increase in amplification obtained with multi-stage radio sets it has become increasingly important to keep the electric and magnetic fields of one stage from inducing voltages in another stage. With an amplification in voltage of, let us say, 1000, it is evident that, if only 0.1 per cent of the output voltage reaches into the input, in the correct phase, self-sustained oscillations will be set up. If the amplifying set is therefore to be kept compact, it is extremely necessary to shield each stage of the amplifier from the others.

The general subject of shielding involves the question of shielding against the steady magnetic fields of permanent magnets and wires carrying continuous current, against the electric fields set up by stationary electric charges, and most important in radio work, against changing magnetic fields produced by alternating or pulsating currents, and changing electric fields produced by moving electric charges, such as the charges on condenser plates connecting to an a.c. supply.

These changing fields, both electric and magnetic, produce increasingly more trouble as their frequency increases, but fortunately the shielding means become more effective at the higher frequency.

Against alternating magnetic fields of low frequency (in the audible range) the device to be shielded should be surrounded by one or more boxes of permeable sheet iron; the thicker the iron, and the higher its permeability, the better is the shielding. Whatever joints there are in the iron box must be well made, so that the reluctance of the joint is small. For very thorough shielding it may be necessary to use two or three sheet-iron boxes, one inside the other. There should be an appreciable air gap between the different boxes. The idea involved in this shielding scheme becomes evident with the help of Fig. 183. The coil M carrying current sets up a magnetic field in the neighborhood of the coil, and we wish to shield space A from this field. By putting the iron box B around the space to be shielded, the flux in space A is reduced to a small fraction of what it is without the box. The flux in space A, traversing a Faraday tube connecting points C

and D, for example, is equal to the difference in magnetic potential between these two points, divided by the reluctance of the tube. Now the difference in magnetic potential is equal to the flux through the iron box, from C to D, multiplied by the reluctance of the box between these two points. As the reluctance of the box is extremely low, there can be but little magnetic potential between C and D, so that the flux between these two points, inside the box, is very small.

Such a box must generally be made in two parts, for ease of assembly of apparatus. The joint thus required will greatly increase the reluctance of the iron path; unless the joint is properly placed it will seriously diminish the shielding action. In Fig. 184 the upper diagram shows the wrong way to joint the box; much flux will traverse the space inside the box because of the magnetic potential required to send the flux across the joint. With the arrangement given in the lower diagram of Fig. 184 no flux has to traverse the joint, and the flux inside the box is only a small fraction of what it is with the joint across the box as in the upper diagram.

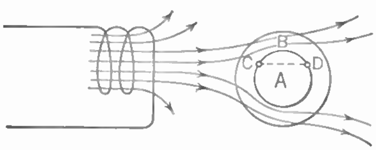


Fig. 183.

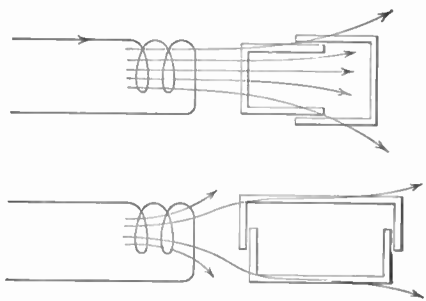


Fig. 184.

When the frequency is as high as perhaps one kilocycle, a copper box is about as efficient a shield as an iron box; for higher frequencies the copper box is better than the iron one, for the same thickness of material. The shielding action of the copper is due to eddy currents in the copper; it can have no shielding action as analyzed above for the iron, because the magnetic permeability of copper is the same as that of air.

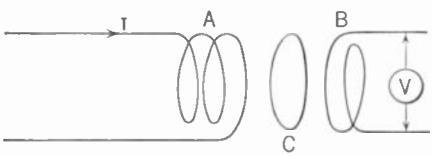


Fig. 185.

The action of eddy currents in shielding against changing magnetic fields will first be analyzed from the arrangement shown in Fig. 185. If an alternating current flows in coil A an alternating magnetic field will be set up in the surrounding region. Suppose for the moment that coil C is absent. A voltmeter connected to the terminals of coil B will give a reading showing that the changing magnetic field from A is generating a voltage in this coil.

Now suppose coil C present, short-circuited, and having a negligible resistance. The changing flux from A will generate a voltage in C, and so current will flow in this coil. The flux through coil B will now be the resultant of that set up by the m.m.f.'s of both coils A and C, and under certain conditions, this resultant will be zero. In this case, the coil C is acting as a perfect shield between coils A and B.

The shielding action of coil C is more perfect as its resistance is decreased, for the following reason. The voltage induced in C is 90° behind the flux from coil A, as we know this voltage is proportional to the *rate of change* of the flux through C. Now as the flux in A is in phase with the current in A (no iron being present) it follows that the voltage induced in C is 90° behind the current in coil A.

Now if the resistance of coil C is negligible (compared to its reactance) the current in C will lag 90° behind the voltage, and so will be 180° behind the current in coil A. Hence the magnetic actions of coils A and C are exactly opposite in phase, and if the placing, and size, of coil C is suitable, the net m.m.f. acting on coil B is zero.

If, however, the resistance of coil C is gradually increased (other things being the same) the current in C will decrease in magnitude, and will come more nearly into phase with the voltage in this coil. Both of these actions diminish the shielding action of coil C, so that if the reading of the voltmeter (Fig. 185) is taken as the resistance of coil C is increased, the readings follow the curve given in Fig. 186. With very large resistance in coil C the reading of the voltmeter on coil B is the same as it is with coil C absent.

A mathematical analysis yields the relation that*

$$\frac{\Delta M_{A-B}}{M_{A-B}} = \frac{K_{AC} \times K_{BC}}{K_{AB}} \frac{\overline{\omega L_c}^2}{Z_c^2}. \qquad (79)$$

in which

M_{AB} = mutual induction of A–B, with C absent
ΔM_{AB} = change in M_{AB}, caused by action of C
K_{AC} = coefficient of coupling of coils A and C
K_{BC} = coefficient of coupling of coils B and C
K_{AB} = coefficient of coupling of coils A and B
ωL_C = reactance of coil C
Z_C = impedance of coil C

* See article on shielding in Proc. I.R.E. for Aug. 1925, by Morecroft and Turner.

When this relation is tested experimentally M and ΔM can be measured in an a.c. bridge (Exp. 10) or the readings of a voltmeter connected to coil B may be used, provided the current in coil A is held fixed. Obviously the resistance of the voltmeter must be high compared to the impedance of coil B, if its readings are to be taken as the voltage *generated* in this coil.

Suppose that the voltmeter reads 3 volts with a certain current and frequency in coil A, coil C being open-circuited. Then with coil C circuit closed, same current and frequency in A, the voltmeter connected to coil B reads 2 volts. The coefficient of shielding is then given by $(3-2)/3 = 33\frac{1}{3}$ per cent.

The three coupling coefficients of eq. (79) can be determined by bridge measurement, if they are not known, as can L_c. By taking a

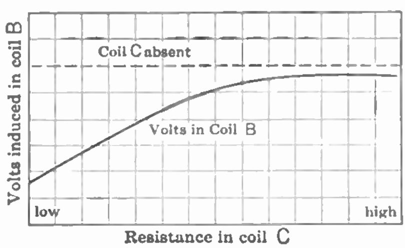


FIG. 186.

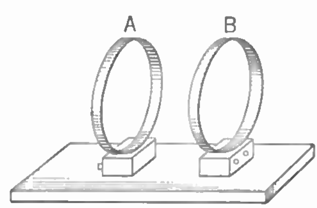


FIG. 187.

a series of readings, with various resistances in the circuit of C, the accuracy of eq. (79) can be tested.

In case a sheet of copper is used between the two coils A and B, in place of coil C, the eddy currents in the copper sheet will have the same kind of action as did the current in coil C. The perfection of shielding depends upon the thickness of the copper, increasing as the thickness of the copper is increased.

Shielding action due to eddy currents is always accompanied by increase in resistance of both circuits; it is generally preferable not to have the copper sheet (or copper cans as are now extensively used around the coils of radio frequency amplifiers) close to either circuit. If it is placed close to either coil (of Fig. 185) the resistance of that coil is unduly increased. Of course the larger the shielding can in which the coil is placed, the greater the cost—manufacturers find that a separation between coil and can of from 0.5 to 1 inch is satisfactory. Only a slight increase in resistance occurs.

From what has been previously said it appears that a shielding container should have low resistance; a high-resistance metal is ineffective, as is a thin can of low-resistance metal.

To investigate the effect of thickness on shielding it is convenient to arrange two short solenoids, say about 4 inches diameter, co-axially, about 1 centimeter apart. Various thicknesses, of various metals, can then be inserted between the coils. A voltmeter reading, with and without the shielding plate in place, gives the shielding coefficient. The metal sheets for the above mentioned pair of coils should be about 6 inches square; if they are smaller than this the shielding will be less than the proper amount.

Of course the shielding can also be measured by an a.c. bridge measurement, M being measured as outlined in Exp. 10, both with and without the shielding present. If this method is used, it is easy to measure the effect of the shielding upon the resistance of either coil; the bridge balance will give resistance as well as inductance. It will be found, in the course of this experiment, that whereas the shielding coefficient continually increases as the thickness of metal increases, the resistance due to shielding is a maximum for a certain thickness of metal. This effect could have been predicted from the results of Exp. 4, in which the resistance of a coil was measured as the resistance in series with another coil (coupled to the one under test) was varied. A certain secondary resistance results in a maximum increase in resistance of the first coil.

A convenient set-up for this test is shown in Fig. 187. The two coils A and B can be multilayer coils, about $\frac{1}{2}$ inch long, 4 inches in diameter, and enough turns of No. 24 wire to give about 10 millihenry inductance. They are supported on blocks about 1 inch high, fastened on the base board about $\frac{1}{2}$ inch apart. A third coil can then be slipped in between these two, and held co-axial with them. By putting various resistances in series with this third coil the relations of eq. (79) can be tested. A high-resistance, low range voltmeter is required to read the voltage induced in the second coil; either a triode voltmeter, or copper oxide rectifier voltmeter will serve.

Various thickness of sheets of different metals, about 6 inches square, can be put in place of the third coil, to measure the shielding effect so obtained. By having some duplicate sheets, some of them having had slits cut in various directions, the effect of bad joints in the shielding containers can be judged.

The 10 millihenry coils specified above are suitable for measurements in the audible range, but if measurements are to be extended to the radio frequency range, three or more sets of coils will be required. Another pair, same size as the first, with only 0.001 henry inductance, will serve to about 1000 kc., and another pair of about 100 microhenrys will serve in the higher radio frequency range.

If time permits, measure the inductance and resistance (at radio frequencies) of a coil similar to those used in the R. F. amplifier of a modern broadcast receiver, both with and without its copper shielding can. The determination is to be made as in Exp. 29.

If time permits, obtain a series of observations to show specifically the comparative merits of iron and copper as shielding material, throughout the audio frequency range.

Plot results about as shown in paper on shielding in the *Proceedings of the Institute of Radio Engineers* (page 477) for August, 1925, and compare your values with those given in that paper.

EXPERIMENT 39

Object.—Study of cathode ray oscillograph and its use in making some radio, and audio, frequency measurements.

Analysis.—The Duddell type of oscillograph, in which a small oscillating mirror makes a photographic trace on a moving film, should be familiar to students before entering the radio frequency laboratory; it may be of much service in making performance tests of the audio frequency, and power supply units of radio apparatus. It is good for frequencies up to about 1000 cycles a second, but requires about 0.1 ampere to give full scale deflection, so will not operate with the small amount of power available in many circuits.

The Einthoven string galvanometer is good for making distortion, and similar, tests of A. F. amplifiers; it requires only about 1 microampere to give a good deflection. However, it is not easy to manipulate, and must be calibrated for each frequency measured. Its sensitivity varies almost inversely with the frequency; a 100-cycle wave, having a twentieth harmonic, as large as the fundamental component, would show a twentieth harmonic ripple only about 7 per cent of the fundamental. Thus a photograph of a distorted wave obtained with this type of oscillograph must be analyzed into its components, and then each component must be multiplied by a proper calibration factor (almost proportional to the frequency), and the various components so obtained are used to synthesize the wave which was photographed.

Both the Duddell type, and the Einthoven string, oscillographs will photograph a non-periodic wave, if the changes are not too rapid.

A cathode ray oscillograph utilizes a beam of electrons as the moving part which records the wave form. In some forms this type of oscillograph utilizes a phosphorescent screen upon which the beam of electrons impinges, to give the picture of the wave form; in other forms, the beam of electrons impinges directly upon the photographic plate upon which the record is to be obtained. By this means a record of a non-periodic wave can be obtained, even if this has a frequency as high as 100 megacycles. However, this type of instrument (known as the Dufour oscillograph) is very expensive and difficult to operate, so the average laboratory will not have it.

In the type of oscillograph to be examined and used in this experiment, a beam of electrons, evaporated from a minute, hot, oxide-coated cathode is accelerated by a plate potential of about 300 volts, and impinges upon a screen of wolframite, inside the glass tube, on its nearly flat end. A luminescent spot on the screen shows where the beam strikes the screen. The beam, on its way from the cathode to the screen, has to pass between two pairs of deflecting plates, one after the other. The two sets of plates are arranged at right angles to each other, so that the second pair urge the electrons at right angles to the direction in which they are urged by the first pair. By impressing voltage waves on these two pairs of plates, the luminescent spot is made to trace figures of various shapes, and from the form of these figures the composition of the voltage waves impressed upon the deflecting plates is deduced.

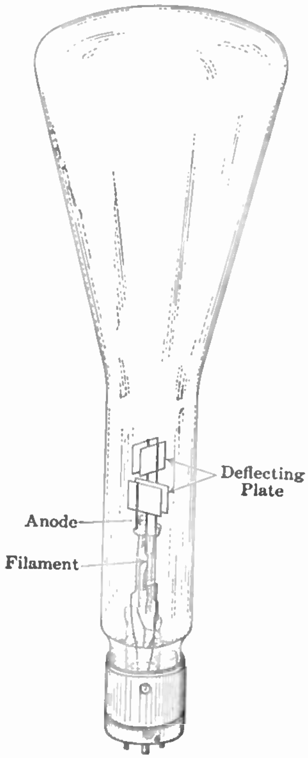


FIG. 188.

The specific form of this cathode ray tube which is readily available to laboratories is made by the Bell Telephone Laboratories (Fig. 188), and the specific information and instruction given below have to do with this tube. (Type 224-A Vacuum Tube.)

The beam of electrons originates on a small oxide-coated filament, which requires from 1.2 to 1.7 amperes and has about 1-ohm resistance (hot). As the filament current must be adjusted in fine steps for getting the proper operating conditions, it is well to use a rheostat of about 4 ohms resistance in series with the filament, and employ a 6-volt storage battery. The plate potential, of from 250 to 400 volts, is best obtained from a tray of *B* batteries. The normal current in the plate circuit is less than 0.001 ampere, so there is no appreciable drain on the plate battery. A resistance of about 2000 ohms, *must* be used in series with the plate battery to decrease the chance of excessive anode currents ruining the tube.

This tube has an appreciable amount of gas left in it, which serves the two purposes of helping to focus the beam of electrons sharply on the fluorescent screen and of preventing charges accumulating on the inside walls of the tube.

The original cathode ray oscillograph (called a Braun tube) contained very little gas, and required from 10 to 50 kv. anode potential. Charges accumulating on the inner walls of the tube had a deflecting action on the cathode beam which made its path in the tube erratic and difficult to control. It was also troublesome to focus the spot, in this low gas pressure tube.

The tube to be used in this test has sufficient leakage (owing to its gas, which becomes ionized when the tube is used) so that charges impinging on the inside walls readily leak back to the anode and furthermore the spot focuses quite sharply, for a given plate voltage, by merely adjusting the filament current to its proper value. This value is different for different plate voltages.

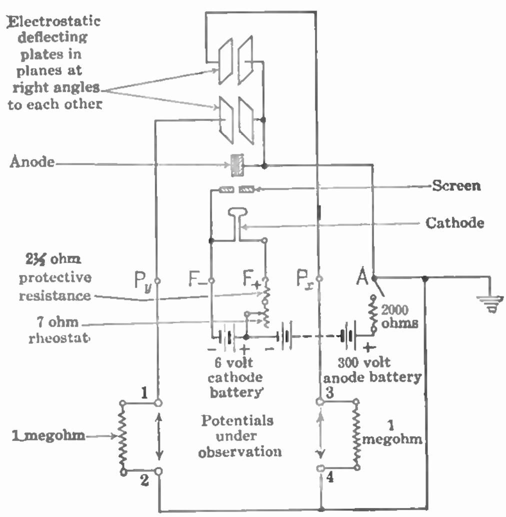


Fig. 189.

The two sets of deflecting plates, at right angles to each other, have one plate of each connected together and to the tubular anode (Fig. 189). As it is generally advisable to ground the anode, the battery heating the filament must be well insulated from ground: it is normally about 300 volts below ground potential.

The deflection of the beam by the plates, with 300 volts between filament and anode, is about 1-cm. movement of the spot on the screen for 10 volts between plates. There is, of course, a slight leakage of current between these plates, because of the ionized gas present in the tube, but it is generally only a few microamperes.

The cathode ray can, of course, be deflected by a magnetic field, as well as by an electric field. The deflection is in the *same* direction as the electric field, and at right angles to the direction of the magnetic field. A permanent magnet brought near the tube may result in such a large deflection that the spot moves completely off the screen.

This deflection of the beam by a magnetic field makes it possible to move the beam by the *current* in a circuit, whereas the deflecting plates operate in accordance with the voltage in a circuit. By putting small coils close to the tube (one on each side) at about the place where the deflecting plates are, reasonable deflection of the spot can be brought about very easily.

Using coils about 5 cm. in diameter, placed co-axially close against the outside of the tube as in Fig. 190 (this makes them about 5 cm. apart), the spot is deflected about 1 cm. for 10 ampere turns. Thus if each coil has 100 turns, a current of 50 milliamperes will deflect the spot about 1 cm. The voltage deflecting plates, inside the tube, are made of high-resistance material, so that the eddy currents set up on them will not interfere with the deflecting action of alternating magnetic fields.

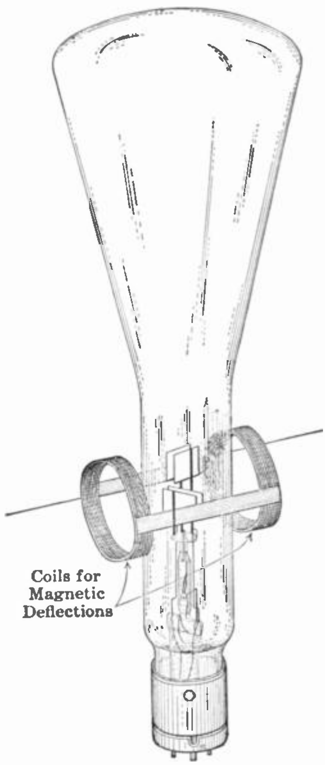


Fig. 190.

The coils should be electrically connected to the anode of the tube so that they assume anode potential. Otherwise they may exert a deflecting force due to their electric field, as well as their magnetic field.

Neither electric nor magnetic deflections are exactly proportional to the deflecting force because of the curvature of the screen. For a deflection of 4 cm. the curvature results in an error of 1.6 mm., or about 4 per cent.

It is advisable to have a small switch in the anode circuit, and impress plate voltage on the tube only during the time it is in use; this will increase the life of the filament, which is normally about 200 working hours.

Uses of the Tube

(1) The tube may be used as a high impedance voltmeter. By pasting a piece of translucent cross-section paper on the end of the tube, and putting the voltage to be measured on one pair of deflecting plates, the measured motion of the spot serves as a voltage indication.

The resistance between the deflecting plates is a megohm or more, so that the voltage to be measured is not seriously disturbed by connecting the plates to the circuit. There is no appreciable frequency error until the frequency reaches several megacycles, so the tube may be used for measuring R. F. voltages. Of course the width of the beam (the spot becomes a beam when alternating voltage is put on one pair of deflecting plates) cannot be measured to better than possibly 5 per cent, but in many measurements this is good enough.

It may be used, for example, to measure the voltage ratio of an A. F. transformer; the secondary circuit of such a transformer has such a high impedance that the use of any voltmeter of less than about 1 megohm resistance seriously decreases the measured value of voltage. This use, of course, enables the response of an A. F. amplifier to be measured throughout the audio frequency range and thus get its frequency-amplification curve.

It may be used to get the hysteresis loop of an iron sample, this being excited by alternating current. If some flux from the iron core to be tested is properly introduced into the tube (in the region of the deflecting plates) the cathode beam may be made to oscillate in a plane parallel to one set of deflecting plates. A resistance is placed in series with the coil which is magnetizing the iron, the value of the resistance being sufficient to produce an RI drop (I being the exciting current) equal to about 10 volts. This RI drop is used to excite that pair of deflecting plates which moves the beam at right angles to the direction in which the magnetic field moves it.

When both the flux from the iron core, and the RI drop, are thus acting on the beam simultaneously, it will trace on the fluorescent screen the rectangular coordinate relationship of flux and magnetizing current. This is, of course, the hysteresis loop of the iron.

In case it is desired to get the form of some function with respect to *time*, one pair of deflecting plates must be excited by a voltage which changes uniformly with respect to time. A portion of the charge or discharge curve of a condenser gives such a voltage. If, for example, a 100-microfarad condenser is connected to a 100-volt line, through 1 megohm of resistance, the voltage across the condenser will rise from zero almost linearly for several seconds. The current which flows when the connection is made is only 10^{-4} amperes, and it will take 10 seconds for such a current to charge the condenser to 10 volts. If then one pair of deflecting plates is connected across this condenser the beam will move (with almost uniform speed) across the screen in 10 seconds. The other pair of plates may be connected to some voltage form which it is desired to view.

If the 100-microfarad condenser is decreased to 1 microfarad the beam will sweep across the screen in only 0.1 second. By arranging the condenser to be charged periodically (and discharged of course in the intervening intervals) by a rotating switch, and by having the same switch set up some transient condition which it is desired to view, the shape of the transient will appear on the screen as a stationary curve.

If a triode is excited by an alternating voltage (say a sine wave) we normally expect the plate current to undergo sinusoidal variations. By putting the grid voltage on one pair of plates, and exciting the other pair of plates by an *R I* drop in the plate circuit, the beam will trace some well-known figure on the screen. If the voltages on the two pairs of plates are equal in magnitude and in phase, the figure will be a straight line in a 45° direction. (It may be necessary to excite the grid of the triode from a potentiometer and use a greater drop, from the same potentiometer, to excite the oscillograph plates, to make the voltage on both sets of plates equal, and of about 10 volts value.)

In case the plate current is out of phase with the grid voltage the beam will trace some sort of figure on the screen. If both voltages (the grid voltage and *R I* drop due to plate current) are sinusoidal the figure will be some sort of ellipse. If the plate circuit is highly reactive so that the plate current lags practically 90° behind the grid voltage the ellipse will be a circle.

In case the grid voltage is sinusoidal and the plate current distorted, the figure will be more complex, departing from the circular, or elliptical, form more as the plate current becomes more distorted.

The oscillograph serves well to investigate a modulated radio frequency current. This current (by means of an *R I* drop) is used to excite one pair of plates, and some uniformly varying voltage (such as the condenser charge analyzed above) is used to sweep the beam across the screen. The separate radio frequency cycles will not be visible, but the envelope of the R. F. current will be visible, and from the form of this envelope the degree of modulation is once seen.

One of the most important uses of this cathode ray oscillograph lies in the field of frequency standardization.

If a voltage of 100 cycles is put on one pair of deflecting plates and a voltage of 200 cycles is put on the other pair, a double-looped figure will appear on the screen, about the shape of a figure eight. If the one frequency is not exactly double the other, the figure will continually change its shape, going periodically through a whole cycle of different forms. Thus if one frequency is variable, it may be adjusted to a value exactly twice that of the other; it is only necessary to adjust conditions until the double-looped figure remains fixed in form.

Now if one frequency is made exactly three times the other a three-loop figure will appear, and if one frequency is ten times the other a figure with ten loops will appear (Fig. 191).

Thus if we adjust some frequency, such as a tuning-fork, or revolving motor, to a very exact low value, other frequencies can be adjusted to exact multiples of this, and can be set with the same accuracy as that attainable with the low frequency. By using say a 1000-cycle tuning-fork as our known fundamental frequency we can adjust an oscillating triode circuit to give us exactly 10,000 cycles (or say 20,000 cycles if the 20 loops can easily be counted).

Now on another cathode ray oscillograph the 10,000 cycles is used as the low frequency and another triode circuit is adjusted to give exactly 100,000 cycles. This frequency is used as the low frequency on another oscillograph on the other plates of which another oscillat-

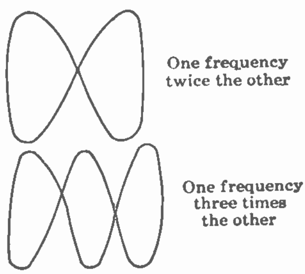


Fig. 191.

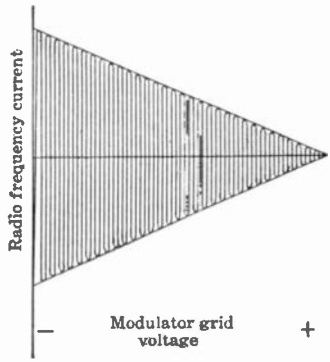


Fig. 192.

ing triode circuit impresses 1,000,000 cycles. By adjusting conditions so that all three oscillographs show stationary figures, the 1,000,000-cycle frequency is known to the same precision as is obtained with the 1000-cycle fork.

A special application of the oscillograph has been made in the adjustment of the so-called Heising scheme of modulation. In this scheme the low-frequency potential impressed on the grid of the modulator is supposed to produce corresponding changes in the amplitude of the radio frequency current generated by the oscillator.

One set of deflecting plates is excited by a drop proportional to the radio frequency current (an RI or XI drop), and the other set of plates is excited from the voltage in the grid of the modulator tube. The figure should be a triangle, as shown in Fig. 192. Here the modulator grid voltage deflects the beam horizontally and the radio frequency current deflects it vertically.

If the R. F. current is proportional to the modulator grid voltage the triangle will be perfect, for 100 per cent modulation. This means that R. F. current has zero amplitude when the modulator grid has its maximum positive value.

If something less than 100 per cent modulation is desired the oscillograph permits the adjustment to be readily made. Thus by changing conditions in the circuit, making the maximum positive grid potential less than it was before, or if it was not positive at all before (for best results it should not be) by making the minimum negative grid

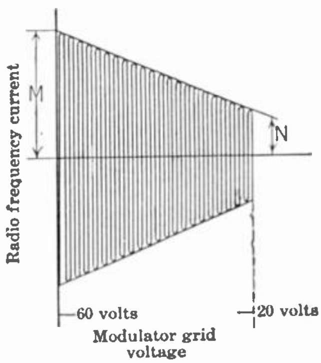


FIG. 193.

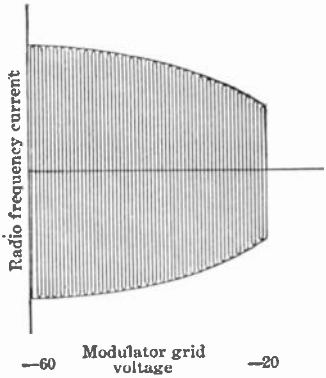


FIG. 194.

potential less than it was before, the percentage of modulation is diminished and the oscillograph gives a figure as shown in Fig. 193. Here the percentage of modulation is given by $100 \times \dfrac{M - N}{M + N}$

In case the modulation is not perfect, the R. F. current does not vary linearly with the potential of this modulator grid. Such a condition results in distortion, the R. F. current carrying modulation frequencies which the modulator should not have put into it. Such an effect is indicated in Fig. 194. This diagram shows that after the modulator reaches a certain negative potential, further potential changes in the same direction do not produce any effect on the amplitude of the R. F. current.

With 300 volts on the plate circuit of the cathode ray tube adjust the filament current to give a well-defined spot on the fluorescent screen. Paste a piece of transparent cross-section paper on the end of the tube. By means of a continuous voltage calibrate this tube as a voltmeter, using first one set of deflecting plates and then the other. If the beam does not lie close to the center of the screen (with no deflecting voltages) bring a bar magnet near the tube and place it in such a position that the spot is brought to its proper place. Leave the magnet in this position while using the tube.

Using the tube as a voltmeter obtain the amplification-frequency curve of a stage of audio frequency amplification. If possible use the same transformer and conditions as were used in Exp. 33.

Arrange the tube to give the hysteresis loop of an iron bar. Trace upon a piece of transparent paper the loop of laminated electric steel, a solid iron bar, and a piece of permanent magnet steel.

Calibrate very accurately one of the laboratory tachometers and by it adjust an alternator to give some exact frequency, say 100 cycles. Use this frequency to excite one pair of deflecting plates. Upon the other pair of deflecting plates put in succession 200, 300, 400, 500, etc., cycles from a calibrated audio frequency oscillator to test the accuracy of its calibration. Notice how high the one frequency can be, with respect to the 100 cycles, and still permit the determination of which multiple it is of the low frequency.

EXPERIMENT 40

Object.—Measurement of amplification, selectivity, and fidelity of a complete radio receiver.

Analysis.—The typical broadcast receiver consists of an antenna circuit, sometimes tuned and sometimes not; two or more stages of tuned radio frequency amplification; a detector; and one or two stages of audio frequency amplification. The last triode of the audio frequency amplifier is called the output tube, as its plate circuit furnishes the power to the loud speaker; it is generally of considerably larger power rating than the other and has a lower plate circuit impedance.

The antenna circuit (if tuned) and the three or more tuned radio frequency circuits preceding the detector give the set its selectivity, or ability to discriminate between the broadcast signals from different stations.

The detector leaves behind the radio frequency, passing along only *the variations in amplitude* of the radio frequency current. These variations constitute the audio frequency current which is amplified by the A. F. amplifier and supplied to the loud speaker. As the signals from all broadcast stations utilize the same A. F. currents, it is evident that all the discrimination between different stations must be carried out before the signal is changed from R. F. to A. F. currents; that is, all of the set's selectivity must come ahead of the detector.

As was investigated in Exp. 22, the triode operates very inefficiently as a detector, if the modulated R. F. voltage impressed upon its input circuit is of low amplitude; for low amplitude signals the amount of rectified current varies with the *square* of the signal impressed upon its input circuit. It is thus the function of the R. F. amplifier to increase the amplitude of the very weak R. F. signal picked up by the antenna, until it has a value upon which the detector can efficiently operate. This generally means a R. F. signal of a few volts amplitude.

If too strong a R. F. signal is impressed upon the input of the detector a distorted A. F. output will result, until special precautions are taken to prevent it. Thus we may say the function of the R. F. amplifier is to increase the amplitude of the weak signal picked up by the antenna to about 1 volt amplitude and to give sufficient selectivity to separate completely signals spaced, say, about 20 kc. apart.

Now a broadcast signal is not of one definite frequency; the modulated R. F. carrier wave is really a whole band of R. F. currents. If the modulation is a result of the microphone picking up an orchestral selection the signal will include practically all the frequencies between that about 6 kc. above the carrier frequency to that 6 kc. below the carrier frequency. Thus a 1000-kc. carrier, modulated by orchestral music, would include all frequencies between 994 and 1006 kc.

All these frequencies should be amplified equally by the R. F. amplifier, but this is far from the fact; those frequencies closest to the carrier frequency are amplified more than those farther away, and these "farther away" frequencies are those which, after detection, give the higher frequencies of the A. F. current. Thus the R. F. amplifier discriminates against the high-frequency components of the modulated wave, and after passing through the detector, the rectifier signal has low-frequency currents of amplitude much greater than they should be, in comparison with the currents of frequencies high up in the audio frequency range.

Thus the R. F. amplifier has greatly increased the strength of the antenna signal, has selected this one frequency (and a narrow band either side of it) for amplification so that it becomes thousands of times as intense as others differing in frequency, but at the same time as it has effected these two necessary tasks, it has discriminated among the various R. F. currents involved in the signal band, so that after passing through the detector the high A. F. currents are less than they should be, in comparison with the low A. F. currents.

The amount of amplification obtained in the R. F. stages of a good receiver is from 1000 to possibly 100,000 in volts; the signal of 10 or 20 microvolts picked up by the antenna appears at the detector input circuit with a strength of a volt or more.

The audio frequency output of the detector is sent through one or two stages of audio frequency amplification before being impressed on the input circuit of the output tube. As was found out in Exp. 33, the A. F. amplifier always discriminates in favor of certain audio frequencies; those frequencies lower than 100 and higher than about 5000 are likely to be amplified less than those between these limits.

It has become the practice to give the sensitivity of a receiving set in terms of the microvolts of signal (30 per cent modulation) required to give an A. F. output of 50 milliwatts. The modulation must of course be sufficient, or the A. F. output of the set will be small. Theoretically, 100 per cent modulation should be used, but such a high degree of modulation is likely to result in distortion of the signal, and so is seldom attempted in commercial broadcasting practice. The figure of

30 per cent specified above is rather less than the value used in the average broadcast transmitter.

The sensitivity of a set is then plotted in terms of the microvolts of a 30 per cent modulated signal necessary to be impressed on the input posts of the set, to give an output to the loud speaker of 50 milliwatts. The modulation frequency is generally taken as 400 cycles for this test.

In Fig. 195 are shown such sensitivity curves for three different receivers; the voltage of the signal input is plotted to a logarithmic scale, as has become the practice in radio literature. It is seldom that a receiver has equal sensitivity over the whole of the broadcast band. The R. F. circuits of receivers *A* and *B* were designed to give their best performance at about 1200 kc.; curve *C* is characteristic of receivers which have a capacity feed back from plate to grid circuits—thus an improperly balanced "neutrodyne" receiver gives this form of sensitivity curve.

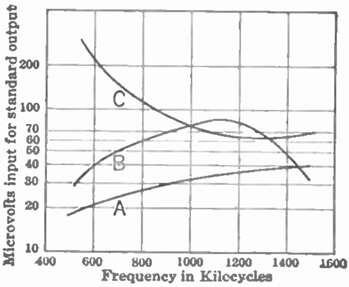


Fig. 195.

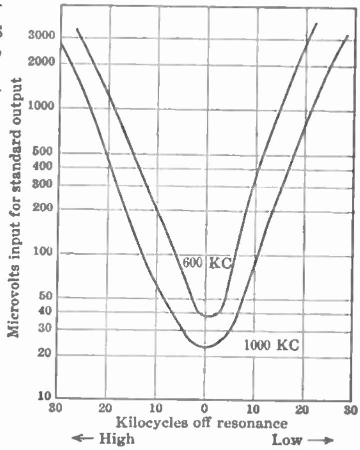


Fig. 196.

The ability of the receiver to discriminate in favor of the desired signal is shown by its selectivity curve. Such a curve is obtained by tuning the R. F. amplifier to the desired signal, and measuring the sensitivity of the receiver for other R. F. signals. Thus with the R. F. amplifier tuned to 1000 kc., and a 1000-kc. signal impressed, the input voltage required to give 50 milliwatts output is measured. Then the frequency of the impressed signal is changed, say, 1 kc., and another sensitivity measurement made, the set still being tuned for 1000 kc. This procedure is repeated for many frequencies either side of that for which the set is tuned; the results look like those given in Fig. 196. The average set has a greater selectivity for stations in the lower part of the broadcast band; the opposite may be true in a set having some regeneration present.

Instead of being sharply selective a set should show equal amplification for frequencies within about plus or minus 6000 cycles of the carrier frequency; the sides of the curve should then rise sharply. Thus in Fig. 197 the full line curve is an ideal selectivity curve, while the dashed line one is the kind ordinarily obtained. The R. F. amplifiers having pairs of weakly coupled tuned circuits in one or more of their stages give a response somewhat as shown in the full line. In Exp. 9 it was shown that the resonance curve of a pair of weakly coupled tuned circuits had the form shown in Fig. 198, and if such circuits are used in

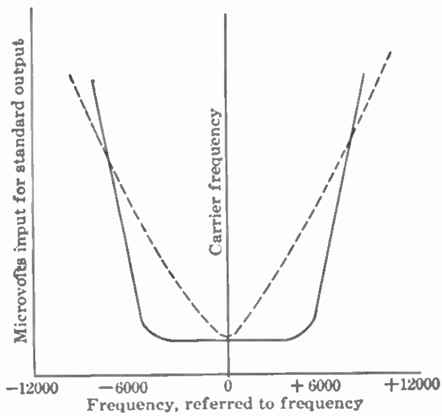


FIG. 197.

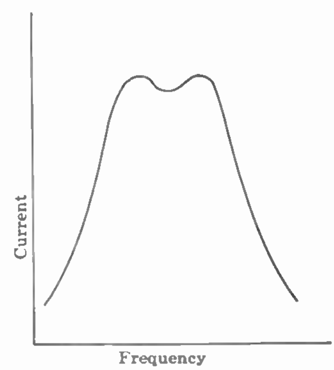


FIG. 198.

the R. F. amplifier they result in a selectivity curve like the full line one of Fig. 198.

The fidelity curve of a receiver shows how it responds to various modulation frequencies. A carrier wave of, say, 30 per cent modulation, is modulated with various audio frequencies, and the A. F. output is measured. For the characteristics previously discussed it has become the custom to adjust the input to get a fixed output, but the fidelity curve is generally obtained in a different way. With a modulation frequency of 400 cycles the input voltage is increased to get standard output. The modulating voltage is then varied in frequency and kept constant in amplitude (so that the percentage of modulation remains fixed) and the power output of the receiver is measured. This power is expressed in terms of the output at 400 cycles, either in percentage, or in decibels.

Such a fidelity characteristic is shown in Fig. 199, the set giving these characteristics is the same as was used in getting the curves of Fig. 196. In this case the output is expressed in percentage of the 400-cycle output, and the modulation frequency is plotted on a logarithmic scale. Such curves should be obtained for at least three carrier frequencies, both at the top and bottom of the broadcast band, and in the middle. In Fig. 196 it is evident that the R. F. amplifier used in these tests is more sharply selective at 600 kc. than at 1000 kc.; it is this greater selectivity of the R. F. amplifier which causes the more rapid falling off in power output at the higher audio frequencies for 600 kc. than for 1000 kc.

To make tests of the kinds analyzed above it is necessary to have a signal generator, which will give signals of various broadcast frequencies, and adjustable amplitude. There must be facilities for modulating this R. F. signal to any percentage desired, at any frequency in the A. F. range. Provision must be made for varying the radio frequency (carrier frequency) by small steps for about 30 kc.

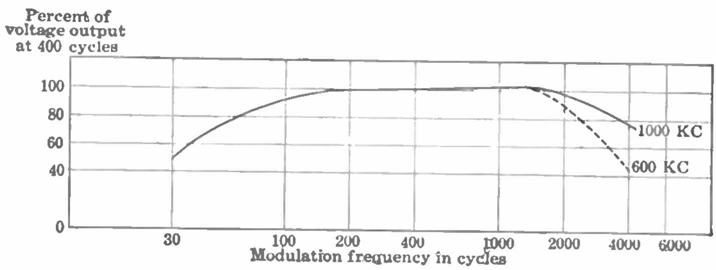


Fig. 199.

either side of the frequency being used. Measuring instruments must be incorporated for reading frequency, percentage of modulation, and microvolts of signal.

It is at once evident that great care must be exercised to completely shield such a generator, or else comparatively large signals will be induced in the receiver by adventitious couplings. When it is remembered that in some cases only 10 microvolts of signal are required, it is evident that the slightest coupling, electric or magnetic, may put into the receiver signals much greater than those intentionally put in by the proper connections. The General Radio Company's Standard Signal Generator is advisable for this test; it has been designed and built specifically for this work. The general outlay of apparatus for making the tests is shown in Fig. 200; all the component parts included in the dotted line are included in the standard signal generator.

In Fig. 201 is shown a connection diagram of the generator with its multi-step attenuator conventionally shown. Referring to this diagram, it is seen that the oscillator is of the "split coil" type, with a variable condenser to get the desired frequency. A third coil serves for getting output; this coil is connected to a fixed resistance, from a sliding contact of which the desired signal is obtained.

A copper plate is arranged to move more or less into the magnetic field of the oscillator, to give the small frequency changes called for in the selectivity test. This adjustment is not shown in Fig. 201.

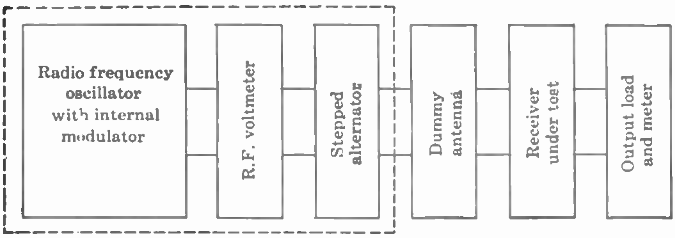


Fig. 200.

An audio frequency oscillator of proper inductance and capacity to give a 400-cycle note is used for modulation. Inspection of Fig. 201 shows that the output of this oscillator is supplied to a transformer, through a potentiometer connection. One coil of this transformer is in series with the plate supply of the R. F. oscillator, thus making the plate voltage of the R. F. triode vary at the 400-cycle frequency by a

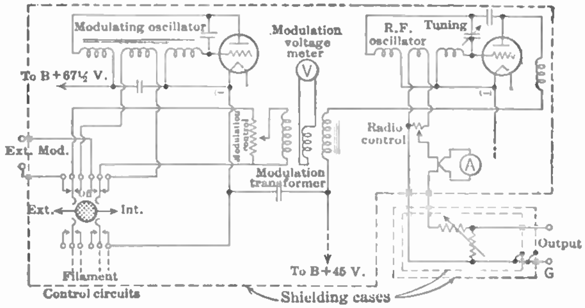


Fig. 201.

controllable amount. A voltmeter (of the copper-oxide rectifier type) serves to show how much of this 400-cycle voltage is supplied. The percentage modulation is obtained from the ratio of the peak value of the 400-cycle voltage in the plate circuit of the R. F. oscillator, to the *B* battery voltage in this circuit. As the modulation voltmeter reads effective values, it is seen that a reading of 10 volts on this meter indicates about 30 per cent modulation when the *B* battery of the R. F. oscillator is giving 45 volts.

A multiple contact switch permits the connection of an outside oscillator to the modulation transformer when desired. This is evidently necessary when getting the fidelity response, requiring frequencies all through the audio frequency range.

In Fig. 201 it may be seen how the very small voltage of the generator output is measured. An attenuation network is supplied with current from the R. F. oscillator through a thermocouple milliammeter. This attenuator has been carefully designed and built to keep its actual attenuation reasonably close to the calculated value, by shielding and proper disposition of its parts.

Instead of being calibrated in milliamperes this thermocouple ammeter is calibrated in microvolts of signal output. The attenuator switch is marked with multiplying factors; by multiplying the reading of the output meter by the factor opposite the attenuator setting, the microvolts of output signal are obtained.

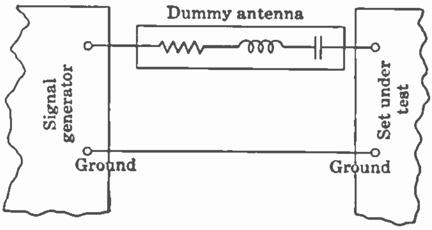


FIG. 202.

The larger changes in signal output are obtained by adjustment of the attenuator and intermediate values by adjustment of the potentiometer between the R. F. oscillator and the attenuator.

As shown in Fig. 200, it is customary to connect the signal generator to the input posts of the receiver under test through a dummy antenna. The proper make-up of this antenna is taken from the rules of the Institute of Radio Engineers as having a resistance of 25 ohms, inductance of 20 microhenrys, and capacity of 200 micro-microfarads. It is connected as shown in Fig. 202.

The output of the receiver under test is best obtained by measuring the current put into a resistive load of the value for which the set is designed. Sets designed for a dynamic speaker should have the speaker replaced by a proper resistance (about 10 ohms for the average dynamic speaker) and those designed for cone speakers should be tested with a resistance load of about 4000 ohms. A suitable thermocouple ammeter to read 50 milliwatts in such a resistance has a full scale reading of 4 milliamperes. If the output is to be read by voltmeter across the resistance, a full scale reading of about 15 volts is required. In both cases the power used in operating the meter must be considered in calculating the output of the receiver, as it may be an appreciable fraction of 50 milliwatts.

Using some standard make of receiver, obtain sensitivity, selectivity and fidelity curves for normal operating conditions.

For 400-cycle modulation, 30 per cent modulation, and 1000-kc. carrier frequency, get a curve between signal input and power output, increasing the signal input in small steps until the output reaches the maximum the receiver can deliver. This curve indicates the distortion which may exist because of low power capacity of some unit of the receiver. The power output should be proportional to the signal input, up to a power output perhaps ten times the normal rating of the receiver. If this overload capacity is not present the set will give amplitude distortion, that is, its output will not faithfully follow the variations of signal strength.

EXPERIMENT 41

Object.—Study and calibration of a c.c. amplifier.

Analysis.—There are three kinds of amplifiers in ordinary use, classified according to the type of circuit used to feed the output of one triode to the input of the succeeding triode. The one most commonly used is the transformer repeating amplifier, one stage of which is shown in Fig. 203. As was found from the results of Exp. 33, such an amplifier has a limited frequency range in which it functions properly. At high frequencies the capacities of winding, triodes, connections, etc., act to short-circuit the secondary winding; at a frequency above about 8000 the ordinary transformer repeating amplifier scarcely amplifies at all.

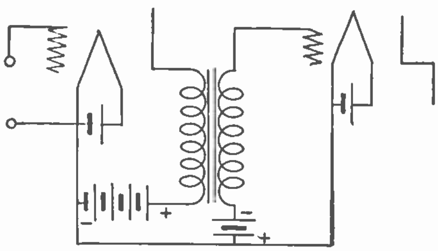


FIG. 203.

At low frequencies the amplifier ceases to function because of the comparatively low reactance of the primary circuit. This reactance should be at least as great as the a.c. resistance of the plate circuit of the triode; but at low frequencies it is much lower than this value. A good A. F. transformer may have 50 henrys of inductance, and the resistance of the plate circuit of the triode may be 8000 ohms. At a frequency of 50 cycles the reactance is about 15,000 ohms, and the amplification is about 90 per cent of its maximum possible value (for this transformer). At 25 cycles the reactance is 7500 ohms, and the amplification is only about 70 per cent of its maximum value; and at 10 cycles per second the reactance has fallen to 3000 ohms, and the amplification is only about 35 per cent of what it should be. For zero frequency, that is continuous current, the amplifier does not function at all.

In Fig. 204 is shown one stage of inductance repeating amplifier; the voltage amplification of this type of circuit has for its maximum possible value the μ of the triode, whereas the transformer circuit (Fig. 203) has as its maximum possible value $n\mu$, where n is the ratio of turns of the transformer.

The upper frequency limit of this type of repeating circuit is somewhat higher than that using a transformer, and its lower useful limit is somewhat lower than that of the transformer, if reasonable values of *L*, *C*, and *R* are employed. Compared to the transformer then, this type of repeating circuit gives about one-third as much voltage amplification, but gives more uniform amplification over a wider range of frequency.

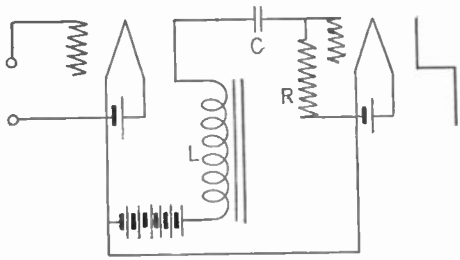


Fig. 204.

The resistance repeating amplifier uses a suitable resistance, instead of the iron core inductance *L* of Fig. 204. As shown in Fig. 205, the drop across a resistance *R* is supplied to the input circuit of the next triode, through a condenser *C*. The resistance *R* serves the same function as it did in Fig. 204, namely, to hold the potential of the grid at some suitable average value, and yet let it fluctuate freely when the signal is being put through. The resistance of *R*, for the arrangements of both Fig. 204 and Fig. 205, must be large (say two or three times as much) compared to the reactance of condenser *C*, at the lowest frequency to be amplified. The value of *R* is generally taken as equal to two or three times the resistance of the plate circuit of the triode; this necessitates the use of a *B* battery with three or four times as much voltage as it required for the inductance and transformer circuits.

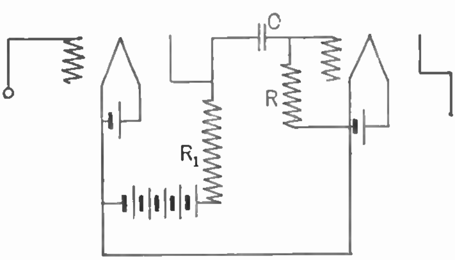


Fig. 205.

The resistance coupled amplifier gives less amplification than either of the others, but its amplification is more nearly uniform, with respect to frequency, than theirs.

Its amplification falls off, at the higher frequencies, for such a frequency that the capacity reactance between the input terminals of the second triode is of the same order of magnitude as the repeating resistance *R*. At the lower frequency limit, the amplification falls off when the reactance of condenser *C* becomes of the same order of magnitude as the grid leak resistance *R*.

A little study shows that all these methods of amplification become inoperative at very low frequencies, say five cycles per second, and that for repeating very slow changes in a c.c. voltage they fail completely. The amplifier to be studied in this test will amplify voltages of any frequency from zero to about 10 kilocycles. It begins to fail at the higher frequencies, because of stray capacities in the circuit.

The arrangement shown in Fig. 206 gives a voltage amplification of about 30, when there is a resistance of 500 ohms in the load circuit. The 350-volt laboratory storage battery is used as a source of power, for both filament and plate circuits. One type 240 triode was used for tube *A*, and for tube *B* four type 171-A triodes were used in parallel. As the four 171-A filaments require more current than does the 240 filament, the latter is by-passed by a 6.5-ohm resistance. The grids of the 171-A tubes are all connected together and to the plate of the 240 tube, and then through 1.2 megohms to the plus terminal of the battery. The filament current, drawn from the 350-volt line, is properly limited by the three resistances (in addition to the filament resistances) of 214 ohms, 125 ohms, and 1.5 ohms. The latter serves to give a small negative bias to the grid of the input circuit, and the other two serve to give the proper grid and plate voltage for the 171-A tubes. Across the load terminals a 500-ohm resistance is normally connected.

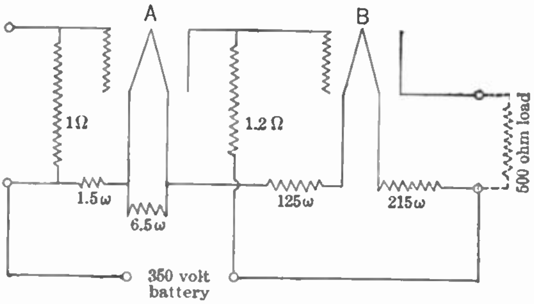


FIG. 206.

It will be seen that a signal on the input circuit affects the plate current of the 240 tube, and this, changing through a resistance of 1.2 megohms, greatly affects the plate potential of the 240 tube and hence the potential of the grids of the 171-A tubes. This in turn affects the plate current through the load, and hence the voltage across the load.

The arrangement shown will result in a current of 0.08 ampere through the 500-ohm load, when no signal is impressed on the input circuit. This gives a voltage across the load of 40 volts. With a signal (c.c.) of 1 volt negative on the input the load current falls to 0.025 ampere, and with 1 volt positive the current increases to 0.145 ampere. The variation is practically linear, and represents, it will be seen, a voltage amplification of $\frac{500 \times (0.145 - 0.025)}{2} = 30.$ The voltage amplification remains at this value until the frequency is increased to 2500 cycles; here it starts to decrease slightly, and at 10,000 cycles it is 19.

A somewhat special method of measuring amplification is necessary, when the frequency is to vary from zero to 10,000 cycles. Filters, to eliminate the continuous current, are in general not suitable for very low frequencies.

Advantage can be taken of the fact that a hot wire meter (or thermocouple) reads the square root of the sum of the squares of the effective values of the separate currents flowing through it. A D'Arsonval type of meter reads only the continuous current flowing through it.

A thermocouple good for 0.100 ampere is put in series with a c.c. meter of 0.100 range, and both of them are put in series with the 500-ohm load. With no signal impressed both meters read 0.080 ampere.

With 1 volt (maximum) signal, of any frequency whatever, there should be flowing in the load circuit, an alternating current of $\frac{30}{500} =$ 0.060 milliampere (maximum) in addition to the steady current of 0.08 ampere.

The c.c meter will then read 0.08 ampere, and the thermocouple will read $\sqrt{0.08^2 + \left(0.06 \times \frac{1}{\sqrt{2}}\right)^2} = 0.091$ ampere. For any frequency, the voltage amplification is then found by first calculating the magnitude of the alternating current, from the readings of the two meters and multiplying this current by 500 ohms.

Thus suppose that with 0.7 volt (effective) impressed on the input circuit at 10,000 cycles, the c.c. meter in the load reads 0.08 and the thermocouple reads 0.088. The alternating current is equal to $\sqrt{0.088^2 - 0.08^2} = 0.0367$ ampere, and the voltage across the load is $500 \times 0.0367 = 18.35$ volts (effective). As the input signal is 0.7 volt the amplification is $18.35/0.7 = 26.2$.

This special amplifier is proportioned to be suitable for amplifying low-frequency voltages for recording on the Duddell oscillograph. The vibration of this instrument requires 0.1 ampere for full scale deflection, so the output of this amplifier is of just the right magnitude to give full scale deflections.

As the input resistance of the amplifier is 1 megohm, it may be put across telephone lines, and similar circuits, to record the form and magnitude of irregular pulses traversing the line. The damping introduced in the line by the amplifier is negligible; events take place in the line in practically the same manner as if the amplifier were not connected.

Construct a c.c. amplifier similar to the one here described and get the relation between load current and input voltage. This curve is to be obtained by c.c. measurements, adjusting the voltage on the input to the desired steady value and reading the load current. Get readings from −1.5 volts to +1.5 volts input, or such other values as vary the load current of the amplifier tested over its complete range.

Get a curve of amplification *vs.* frequency, in the manner described above. Check a few points by measuring the a.c. drop across the load by a sensitive thermocouple, in series with a condenser to block out the continuous current. Thus suppose that in Fig. 207 the drop (a.c.) across the 500-ohm load is 20 volts, and that the thermocouple has 800 ohms resistance, and requires 0.002 ampere for full scale deflection. To keep the alternating current in the meter down to 0.002 ampere, with 20 volts impressed, requires 20,000 ohms resistance.

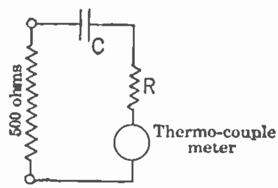


Fig. 207.

Hence 19,200 ohms is introduced at *R*. The condenser *C* must be of such a capacity that its reactance, at the frequency being tested, is small compared to 20,000 ohms.

A 1-microfarad condenser has 16,000 ohms reactance at 10 cycles per second, so that the voltage read by the thermocouple would be about 20 per cent low. If, however, 10 microfarads were used, practically all of the a.c. drop across the 500-ohm load would be measured by the thermocouple. This, in series with the resistance *R*, can be calibrated in volts on a c.c. circuit, such as a *B* battery.

EXPERIMENT 42

Object.—To study the internal capacity of various types of coils, by different methods.

Analysis.—The internal capacity of a coil may have a very serious effect in limiting the frequency range available, in the tuned circuit of which it is a part. Further, in the case of coils forming the secondary circuit of an audio frequency transformer, the internal capacity of the coil is the one factor which limits the range of high frequencies which the circuit will amplify. This experiment will be limited to the study of the internal capacity of coils intended for use in tuned radio frequency amplifiers.

Coils for radio circuits may be built in various ways, single-layer solenoids wound on cylindrical forms of "Bakelite" or similar material, two- or three-layer solenoids with or without sheets of insulation between layers, bank-wound multilayer coils, coils wound on forms so that the wires are for the most part in contact with no solid dielectric. They may be impregnated or not, according to the judgment of the radio designer. The turns may be wound close together, or they may be separated from each other by an appreciable distance. All these different methods of construction result in different internal capacities, and we shall study a few of them in this test.

The simplest coil is a single-layer solenoid, wound on a cylindrical form. If such a coil is connected to a variable condenser, and the capacity of the condenser be varied from maximum to minimum, it will be found that the wave length of the circuit varies only in the ratio of between three and four to one. The internal capacity of the coil, in addition to the zero setting capacity of the condenser, is what limits the shortest wave length available.

The internal capacity of the solenoid is due to the wire surface on one end of the coil acting as a condenser plate, the other condenser plate being the other end of the coil. In addition to this "gross" capacity, each turn of wire acts with its adjacent turns to form small condensers distributed throughout the length of the coil. The integral effect of all the separate condenser actions in the coil is used in defining the internal capacity of the coil.

The experimental method used in getting the value of the internal capacity is based upon a fallacy which results in an error of about 50 per cent, but as all recorded results involve the same error, they serve to give comparative results.

The formula for wave lengths of a resonant circuit is $\lambda = 1884\sqrt{LC}$ in which λ is given in meters, L in microhenrys, and C in microfarads. Squaring this equation, and assuming that L is a constant, gives the relation $\lambda^2 = KC$. If then a series of readings are taken between capacity and wave length, and a curve plotted, it should be a straight line passing through zero, but actually it has the form shown in Fig. 208. It is a straight line in so far as measurements can be obtained, between the C_{max} and C_{min} of the variable condenser used. Extrapolating the straight line backwards it cuts the X axis to the left of the "zero capacity" point, by an amount indicated in Fig. 208 by the capacity OA. This is taken as the internal capacity of the coil.

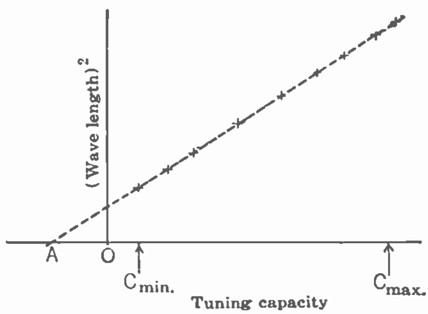


FIG. 208.

Such a determination of internal capacity is evidently based upon the assumption that the self-induction of the coil remains constant, independent of frequency. Now such is far from being the fact. At the longer wave lengths, where the external capacity is large compared to the coil capacity the current throughout the entire length of the coil is essentially the same. For this condition the self-induction of the coil is the same as would be determined by a bridge measurement, at, say, 1000 cycles.

Now when the external tuning capacity is reduced to nearly zero the current in the center turns of the coil is much larger than the current in the end turns. The current in the center turns is sufficient to charge the condenser made up of the internal coil capacity, as well as the small external capacity. The current in the end turns is only that required by the external capacity. Such a diminution of the current in the end turns of the coil results in a corresponding diminution of flux through the coil, and hence a corresponding decrease in the self-induction.

Consideration of this erroneous assumption (constant L) shows that the distributed capacity of the coil determined by this method is considerably *less* than the true value.

Another method of determining the internal capacity of the coil is to measure its natural wave length. The coil is suspended on a string near a triode oscillator, so as to be coupled to it, and the frequency of the oscillator is varied until the slight dip in the current of the oscillator tuned circuit shows that the coil is in resonance with the frequency of the oscillator. A wave meter is then used to determine the oscillator frequency, or wave length.

The inductance of the coil is then determined, by bridge measurement or otherwise. The natural wave length of the coil, and its inductance being known, the internal capacity is calculated from the relation $\lambda = 1884\sqrt{LC}$.

This method of determining capacity is evidently based upon the same erroneous assumption as the foregoing one, namely, that the coil's inductance is independent of frequency.

As a check on either of the foregoing methods the coil may be immersed in some suitable dielectric such as kerosene. The new value of internal capacity should be equal to that previously determined, multiplied by the specific inductive capacity of the kerosene.

If the coil is impregnated with shellac or similar compound, the kerosene evidently cannot penetrate, so that such part of the internal capacity as is situated in the impregnated space cannot increase when the coil is immersed. This will result in an increase in the determined value of internal capacity less than the expected amount.

In order that this experiment may yield valuable results, several coils, of the same inductance, with different construction, should be used; coils of about 1 millihenry inductance are suitable. All of them should be of the same diameter, and wound with the same size wire. Suitable coils are: two single-layer solenoids of D. C. C. wire wound close, one impregnated with shellac and one with no insulating varnish; a single-layer solenoid with spacing between turns equal to about half the diameter of the wire; two two-layer solenoids, one with no paper between layers and the other with two or three layers of paper between; two three-layer solenoids, one with and the other without paper between layers; a three-layer bank-wound coil.

Measure the internal capacity of these coils by the two methods given and check the values for the two un-impregnated single-layer solenoids by immersion in kerosene.

EXPERIMENT 43

Object.—To study the action of filters, of both high pass and low pass types. Characteristics of single sections, and of several sections in series.

Analysis.—An inductance offers an impedance which increases directly with the frequency; a condenser offers an impedance which decreases with frequency increase, and a resistance offers impedance the same for all frequencies. By a proper combination of circuit elements employing these various kinds of impedance it is possible to oppose a high impedance to low frequencies and very low impedance to high ones, or vice versa. A circuit which thus acts selectively

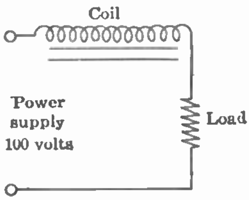


Fig. 209.

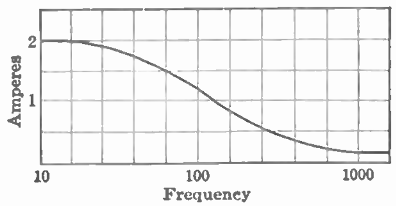


Fig. 210.

towards certain frequencies and opposes others, especially when the arrangement shows this selectivity to an unusual degree, is called a filter.

A simple coil in series with a power supply is really a "low pass filter," that is, it lets through the low frequencies and prevents the high ones from flowing. Thus the arrangement in Fig. 209 is a low pass filter, of very simple type. If the inductance is 0.1 henry and the resistance (total) is 50 ohms, the current which flows in the load for 100 volts impressed at various frequencies is shown in Fig. 210. This then is a low pass filter.

In Fig. 211 is shown a simple high pass filter, and the curve of current *vs.* frequency is shown in Fig. 212. In the arrangements shown in both Fig. 209 and Fig. 211 the change of current, with respect to frequency, is not rapid; it is said that there is not a sharp " cut off." It is possible to increase the sharpness of cut off by using both capacity and inductance in the same filter, properly aranged.

In Fig. 213 there are shown two such filters; that in *a* is a low pass filter and that in *b* is a high pass one. It is easy to calculate the relation between frequency and load current when only one or two sections are used in the filter, but becomes tedious for a filter having many sections, unless a special method of solution is used. Such books as Pierce's "Electric Oscillations and Electric Waves" or Johnson's "Transmission Circuits for Telephone Communication" give theoretical discussions of the multi-section filter problem.

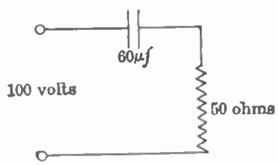


FIG. 211.

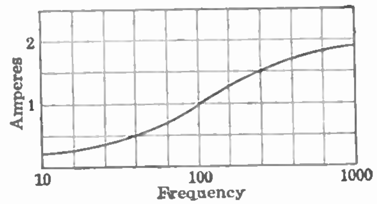


FIG. 212.

In such filters as those shown in Fig. 213 there are two characteristics of the circuit which must be suitably chosen for the purpose intended. The "surge impedance" of the filter Z must be made the same as the load impedance, and the desired "cut-off" frequency must be obtained by using proper values of inductance and capacity.

In Fig. 214 is shown one unit of a high pass filter, having a series capacity equal to C_1 and a transverse inductance L_1; this is called a

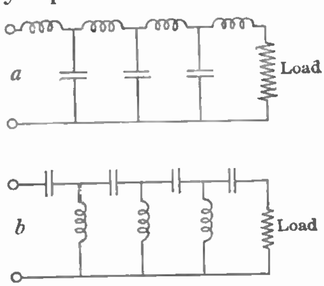


FIG. 213.

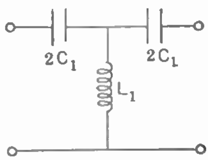


FIG. 214.

T section. It will be noted that there are actually two condensers in series each equal to $2C_1$; because of the series connection the effective capacity of the section is only C_1.

If we wish the "cut-off" frequency to be F_0 it is shown in such books as Pierce (above) that

$$F_0 = \frac{1}{4\pi\sqrt{L_1C_1}}. \qquad (80)$$

and

$$Z_0 = \sqrt{\frac{L_1}{C_1}}. \qquad (81)$$

From these two expressions we derive the approximate relation that, for this filter,

$$C_1 = \frac{0.08}{Z_0F_0} \text{ farads}$$

and

$$L_1 = \frac{0.08Z_0}{F_0} \text{ henrys.}$$

As the proper load resistance of the filter is equal to $\sqrt{\frac{L_1}{C_1}}$, it is seen that at the "cut-off" frequency, $2\pi fL$ is equal to half the load resistance, and $1/2\pi fC$ is equal to twice the load resistance. To increase the selectivity of the circuit, that is, to discriminate more between the high and low frequencies, several sections are generally connected in series, as shown in Fig. 213*b*.

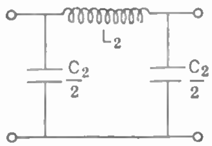


FIG. 215.

The low pass filter of Fig. 213 is really made up of a series of so-called π sections which are connected in series. One of the sections is shown in Fig. 215; the inductance per section is L_2, and there are two condensers each of capacity $C_2/2$ in parallel, making the capacity per section C_2.

For such a section we have relations somewhat similar to those of the T section, namely,

$$Z_0 = \sqrt{\frac{L_2}{C_2}}. \quad \quad (82)$$

and

$$F_0 = \frac{1}{\pi\sqrt{L_2C_2}}. \quad \quad (83)$$

these two conditions yield the equations

$$C_2 = \frac{0.32}{F_0Z_0} \text{ farads and } L_2 = \frac{0.32Z_0}{F_0} \text{ henrys.}$$

In this type of filter, at the "cut-off" frequency, $2\pi fL_2$ is equal to twice the load resistance, and $1/2\pi fC_2$ is equal to half the load resistance.

It is impossible to build coils and condensers without loss, so the above theoretical relations, derived on the assumption of resistanceless coils and perfect condensers, are not quite correct.

In Table I are given the characteristics of a resistanceless high pass filter, of 10 sections, giving the ratio of voltage across the tenth section compared to the voltage in the first section, and in Table II there are corresponding values shown for a line in which the coils have a resistance of 250 ohms per henry.

Table I		Table II	
Impressed F / F_0	$E_{10th\ section}$ / $E_{imp.}$	Impressed F / F_0	$E_{10th\ section}$ / $E_{imp.}$
0.4	0.0	0.4	0.0
0.8	0.0	0.8	0.0
0.9	0.00009	0.9	0.000068
0.96	0.0032	0.95	0.00136
0.99	0.021	1.00	0.044
0.994	0.111	1.05	0.25
0.998	0.273	1.10	0.374
1.0	1.00	1.20	0.525
2.0	1.00	1.40	0.694
4.0	1.00	1.60	0.778
....	....	2.00	0.865
....	....	3.00	0.942
....	....	4.00	0.967

In Fig. 216 are shown the experimentally determined characteristics of one section of L. P. filter with a cut-off frequency of 1000 cycles,

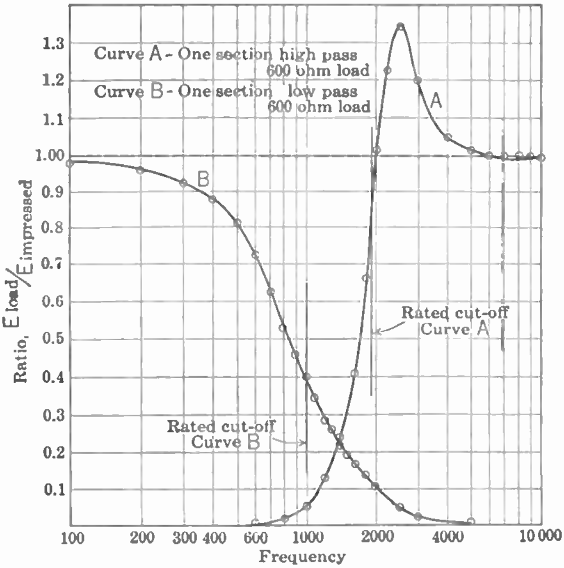


Fig. 216.

and one section of H. P. filter with a cut-off frequency of 2000 cycles. In Fig. 217 are shown the approximate values of inductance and capacity in making up these filter sections. Both of them are built with a characteristic impedance of 600 ohms.

In obtaining the curves given above, a resistive load of 600 ohms was connected across the output terminals. A voltage of adjustable frequency was impressed across the input terminals, and the voltage across the load was read. The ratio of load voltage to input voltage was used as ordinate of the curves of Fig. 216.

Evidently either of these filter sections may show a resonant condi-

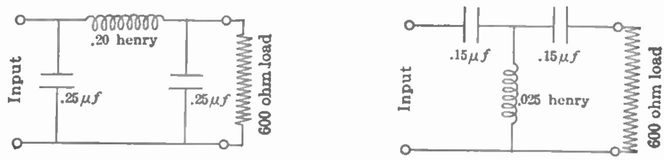


FIG. 217.

tion for some value of impressed frequency. In the H. P. type the condenser of 0.15μf in series with the coil of 0.025 henry will show resonance at the frequency of about 2500 cycles, and inspection of the curve for the H. P. filter does show a resonance peak at just this frequency. The resonance is not as pronounced as it would normally be, because the 0.025-henry coil is shunted by a resistance of 600 ohms in series with the 0.15μf condenser. This shunt path around the coil decreases the sharpness of the resonance phenomenon very much.

It might seem that there should be a resonance peak in the characteristic curve of the low pass filter also, but such is not apparent on the curve sheet, and analysis of the action of the circuit gives the reason.

The resonant frequency for a coil of 0.20 henry in series with 0.25 microfarad is about 700 cycles. At this frequency the reactance of the 0.25μf condenser is about 800 ohms, and shunting this reactance is the load resistance of only 600 ohms. Such a condition practically suppresses any resonance phenomenon.

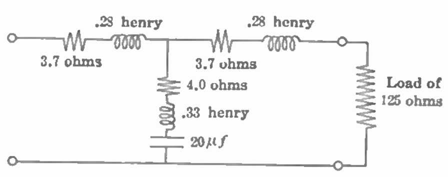


FIG. 218.

By putting several sections of filters (similar to those of Fig. 217) in series the discrimination between the high and low frequencies is accentuated, in fact it is increased in direct proportion to the number of sections used in series.

It is possible to build filters which give a much sharper cut off than those shown in Fig. 217, but in general the filters which give the sharper cut-off are correspondingly more complex.

In Fig. 218 is shown a section of L. P. filter which shows much sharper cut off than that given in Fig. 216. When a load of 125 ohms resistance is connected to its output terminals it performs as shown in Fig. 219, curve *A*. In another section of the same type the line inductances were 0.336 henry and 5.13 ohms resistance, the shunt path had a coil of 0.191 henry and 3.74 ohms, with a 12-μf condenser. When supplying a 125-ohm resistive load this section performed as in curve *B* of Fig. 219. In this same diagram is shown (curve *C*) the performance of a section made up as shown in Fig. 220. This is a high pass type of filter, having the cut-off frequency accentuated by the series resonant shunt path.

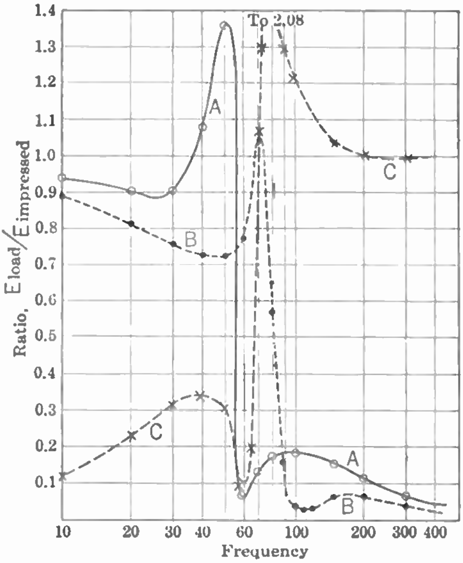


FIG. 219.

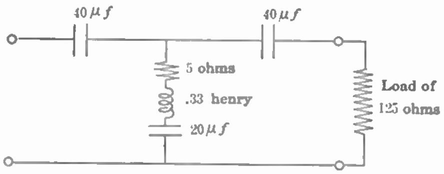


FIG. 220.

Curve *C* of Fig. 219 was obtained from this filter section when a resistive load of 125 ohms was connected across its output terminals.

By using somewhat more complex arrangements it is possible to construct "band pass" filters, or "band rejector" filters. The simplest such scheme is to use a low pass filter in series with a high pass filter, the cut-off frequencies being properly chosen to pass the desired band. Thus suppose a low pass filter with a cut off at 3000 cycles, in series with a high pass one cutting off at 1000 cycles, the characteristic curves being as shown in Fig. 221. The combination filter will have the characteristic

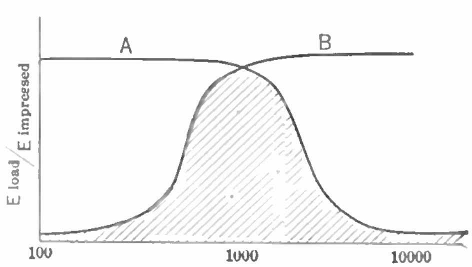


FIG. 221.

surrounding the shaded area, passing frequencies between about 800 and 3000 cycles.

Obtain the characteristics of a low pass filter of one section, and then of three sections in series. Connect a resistive load of the proper resistance to the output terminals; impress some suitable voltage on the input terminals and measure this as well as the voltage across the load. Obtain a series of such measurements over a frequency range from 0.1 to 10 times the cut-off frequency of the filter. Get enough readings, properly spaced with respect to frequency, to plot an accurate curve of its performance.

Carry out the same tests for a high pass filter. Then obtain curves of performance of a band pass filter, by connecting in series suitable H. P. and L. P. sections.

If time permits, get the characteristics of a filter made up of sections similar to those of Figs. 218 and 220 and compare with the performance of the simpler types of Fig. 217.

EXPERIMENT 44

Object.—Study of a simple band pass filter for radio frequency circuits. Effect of varying the amount of coupling.

Analysis.—An ordinary radio broadcast signal utilizes not only the carrier frequency of the station, but a whole band of frequencies on either side of the carrier. If the modulating apparatus of the transmitting set is properly designed, it will impress upon the triode generating the high-frequency current a modulating voltage, which is made up of frequencies from about 40 per second to about 5000 per second. The lower frequencies are found in organ selections, and the higher ones are necessary for the consonant sounds of the voice. When such a modulating voltage acts upon the high-frequency triode, this generates not only its carrier, say 1,000,000 cycles, but in addition two bands of frequencies on either side of the carrier. This idea is demonstrated in Fig. 222, in which the frequencies are not shown quite in proper proportion. The narrow separation between the carrier and the two frequencies 999,960 and 1,000,040 cannot be shown by any ordinary tuning instrument, so we may say that the broadcast signal utilizes all the frequencies between 994,000 and 1,006,000.

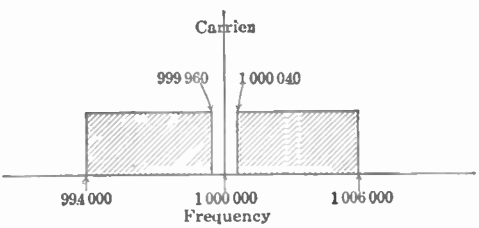


Fig. 222.

Now all the frequencies in this band should be amplified equally by the tuned radio frequency amplifier, but such is seldom the fact. The shape of the resonance curve of an ordinary R. F. tuned amplifier is the same form as the curve obtained in Exp. 6, reproduced in Fig. 223. The circuit is tuned for the carrier frequency, namely, 1,000,000 cycles, and it then amplifies other frequencies according to the values shown in Fig. 223. Thus those frequencies of the two side bands which are close to the carrier frequency are amplified two or three times as much as the outer frequencies of the side bands, that is, the 994 and 1006 kc. frequencies.

Now these outer frequencies of the side bands are those which, after rectification by the detector, give the higher audio frequencies of the program, namely, the consonants of the voice and the very high notes of such instruments as the flute and piccolo.

It is then evident that the ordinary R. F. tuned amplifier has frequency discrimination, in favor of the lower notes of the audio frequency range. It so happens that the average A. F. amplifier tends to correct this fault of the R. F. amplifier, by not amplifying the lowest notes of the A. F. signal as much as it does the high ones. However, it is advisable to correct the defect of the R. F. amplifier if possible, and this can be done to a considerable extent by utilizing the action of two tuned coupled circuits, as brought out in Exp. 9.

Two tuned circuits coupled together show a resonance curve with two peaks; the separation of these peaks depends upon the coupling,

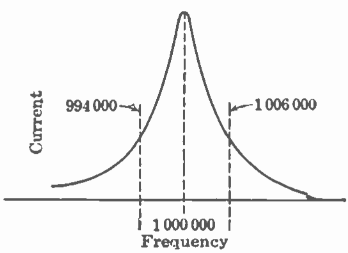


Fig. 223.

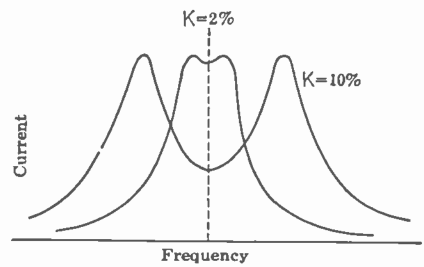


Fig. 224.

the tighter the coupling the greater their separation. In Fig. 224 are shown two typical curves for couplings of perhaps 10 per cent and 2 per cent. The separation of the two peaks, in cycles, is equal to the carrier frequency multiplied by the coefficient of coupling.

If then we want the R. F. amplifier to show equal amplification from 994 kc. to 1005 kc. the coupling should be $\frac{1006 - 994}{1000} = 1.2$ per cent. Now it becomes evident, upon some reflection, that this scheme cannot work equally well throughout the range of broadcast frequencies. Thus at 600 kc. carrier a 1.2 per cent coupling gives a width to the resonance curve (separation of the two peaks) of only 1.2 per cent $\times$ 600 = 7.2 kc.; we want it to be 12 kc. And at the upper end of the broadcast range, say at 1500 kc., the same coupling gives a peak separation of 1.2 per cent $\times$ 1500 kc. = 18 kc. This is not only wider than necessary but it is wide enough to cause interference from other stations in the same part of the broadcast spectrum.

To keep the width of the resonance curve, the same number of kilocycles width it is necessary to use only one-third as much coupling for the higher frequency as is used for the lower.

One circuit arrangement which has been used to utilize the coupled circuit resonance curve is shown in Fig. 225. The two coils, L_1 and L_2, are of about 250μh each, and are arranged to have no mutual inductance. The small coil L_3, of about 3μh inductance, is common to both tuned circuits, L_1–C_1, and L_2–C_2. The untuned antenna circuit is loosely coupled through coil L_4 to coil L_1.

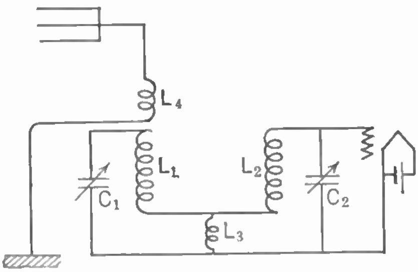


FIG. 225.

The per cent coupling for the two tuned circuits is given by the ratio

$$\left(\frac{L_3}{L_1 + L_3}\right) \times 100.$$

Arrange a circuit as shown in Fig. 226, the two coils L_1 and L_2 being so placed that there is no mutual induction between them. The two condensers C_1 and C_2 should be identical, and are preferably on the same shift, for common control.

Power is supplied to the circuit through a small coil L_4, coupled to coil L_1. An ammeter A serves to check the constancy of the cur-

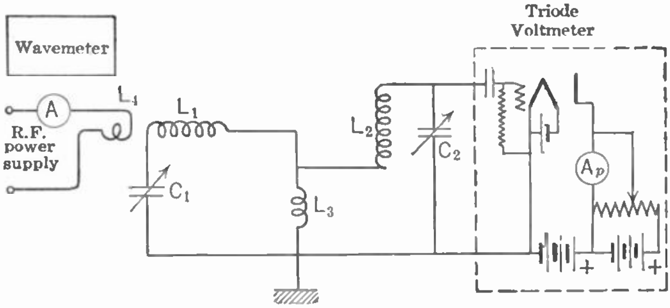


FIG. 226.

rent in L_4 while a test is being made. The voltage across condenser C_2 is read by a vacuum tube voltmeter, of the grid rectification type; its readings serve as a measure of the voltage across C_2.

With a value of L_3 about 1 per cent of L_1, the power supply is adjusted to give a frequency at the lower end of the broadcast band, and the L_1–C_1 and L_2–C_2 circuits are tuned to it. The coupling of L_4 and L_1 is then adjusted to give a reading on the triode voltmeter well up on its scale. The current A is noted and kept constant. The frequency of the supply is then varied through a range of frequency sufficient to plot the resonance curve of the coupled circuits, the readings taken being volts across C_2 *vs.* frequency.

The frequency of the supply is then changed to a value towards the top of the broadcast band, and another resonance curve is obtained here, after the L_1–C_1 and L_2–C_2 circuits have been retuned to the frequency of this higher frequency. It will be found that the width of the resonance curve is about three times as many cycles as it was for the first curve obtained.

Repeat these two runs with a new coil for L_3, having about $10 \mu h$ inductance.

Based upon the results obtained in this experiment design a circuit which will keep a constant width of resonance curve, throughout the broadcast frequency band.

EXPERIMENT 45

To study the action of a detector and obtain the relation › frequency voltage available in plate circuit, in terms of . percentage modulation, of the *R.F.* signal impressed upon ․.

Analysis.—The detector stage of an ordinary radio receiver serves as the "frequency converter" between the radio frequency amplifier

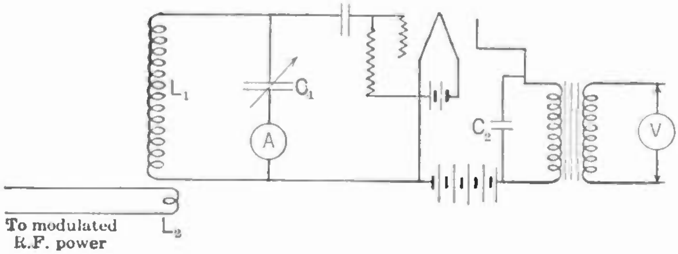


FIG. 227

and the audio frequency amplifier. Its input is connected to the last tuned circuit of the R. F. A., and its output is connected to an iron core transformer which supplies the input to the A. F. A. In Fig. 227 this circuit element has been shown as it is used in the actual set, with the exception that an ammeter *A* is shown in the tuned input circuit, and a voltmeter *V* is shown connected to the secondary circuit of the output transformer *T*. In this test high-frequency modulated power is to be supplied to the input circuit by coupling with coil L_2, which is in turn connected to a suitable source of power. Or the L_1–C_1 circuit may be opened, and the output circuit of a "signal generator" be connected, as shown in Fig. 228.

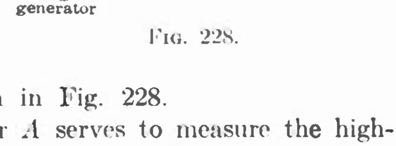


FIG. 228.

A suitable thermocouple ammeter *A* serves to measure the high-frequency voltage impressed on the detector input terminals. In a resonant circuit the condenser reactance must be equal to the coil reactance, hence if the inductance of the coil and the fre signal are known, the coil reactance (hence condenser r easily calculated. This reactance multiplied by the current ter A gives the value of the R. F. voltage impressed on the input terminals.

When using grid condenser rectification the maximum signal a ble in the detector input is generally assumed to be not more tha or 3 volts. More than this results in a distorted audio frequency ou put from the plate circuit. In a circuit such as that of Fig. 228 the "standard signal generator" is able to impress as much as about 10 volts on the detector input. The maximum voltage available from this generator is 0.2 volt, and if the ratio of reactance to resistance of the tuned circuit is 25 (which is a reasonable value for the circuit as arranged here), the voltage on the input circuit of the detector will be 5.0 volts. Of course the actual reactance of any tuned circuit is zero; the reactance referred to is that of either the coil or condenser.

If a higher voltage than 5.0 is desired for the test it will be necessary to use another source of high-frequency power (with adjustable modulation) or else to obtain more voltage on the input circuit by cutting down the resistance of the tuned input circuit. A low-resistance thermocouple ammeter, and a coil having a low high-frequency resistance, must be used if 10 or 15 volts are to be obtained on the input circuit.

It will be possible to use a low-resistance couple for this part of the test because at the higher voltage the circuit will be carrying a higher current, which of course is necessary for a low-resistance couple.

The voltmeter used to read the audio frequency voltage on the secondary of the output transformer must be of very high resistance, several hundred thousand ohms at least. It was shown in Exp. 33 that when even as much as a megohm resistance is connected across the transformer secondary terminals, the voltage between these terminals falls appreciably. Even the vacuum tube voltmeter, which is to be used to read this voltage, will cause an appreciable drop in the secondary voltage from the value it has with free secondary terminals, but, as the vacuum tube voltmeter has about the same resistance as the input circuit of an ordinary amplifying triode, its voltage reading is the value which would be operative in an actual amplifier circuit.

The condenser C_2 of Fig. 227 permits the radio frequency fluctuations to flow in the plate circuit without having to pass through the high impedance of the transformer primary. Without this condenser, the detecting action of the triode is appreciably diminished, unless the primary winding of the transformer itself has sufficient internal capacity to by-pass the high-frequency currents around the primary inductance.

The variation in amplitude of the radio frequency current (that is, its modulation) causes a varying negative charge to accumulate on the grid of the detector triode; higher amplitude of R. F. current means a greater negative charge on the grid, and vice versa.

This varying charge has to leak off the grid in a rather short time, if high-frequency modulation is to be followed; this means that the time constant of the C–R circuit, Fig. 227, must be only about 0.0001 second if the highest voice frequencies are to be properly detected and amplified by the triode.

If a 0.01 microfarad condenser were to be used for C and a resistance of 1 megohm be used for R the time constant would be only 0.01 second, and practically none of the consonant sounds of the voice would appear in the audio frequency output of the detector.

With normal values of C, R, and C_2, find the relation between output voltage with a fixed amplitude of carrier current and fixed modulation, as the carrier frequency is varied over the broadcast range.

With C, R, and C_2 normal, find the relation between A. F. voltage output and R. F. voltage input, the modulation remaining fixed at 50 per cent. Get curves of this relation at the top and bottom frequencies of the broadcast range, and one frequency in the middle.

At some convenient frequency find how the A. F. voltage output varies with the percentage modulation, the amplitude of the carrier current being held constant at about 5 volts.

With the same carrier current frequency and amplitude, and percentage modulation (say 50 per cent) find how the A. F. output voltage varies with the frequency of the modulation voltage, varying this from, say, 50 cycles to 8000 cycles per second. This test is to be carried out with normal conditions in the detector circuit.

Carry out the above run for a value of grid condenser of 0.01 microfarad, with 1 megohm leak resistance.

Find the effect on the detecting action of the triode of removing the by-pass condenser C_2.

Try the effect of varying the grid leak, from several megohms, down to a few thousand, the grid condenser remaining at its normal value.

With conditions normal, say middle R. F. frequency, 5 volts amplitude, 50 per cent modulation, find how A. F. output varies with the voltage used in the plate circuit of the detector.

In some of these tests it may well happen that the A. F. voltage output may exceed the range of the vacuum tube voltmeter. In such case put across the secondary terminals of the output transformer a million ohms made up of, say, 300,000 ohms and 700,000 ohms in series and connected to the V. T. voltmeter across the 300,000 ohms. With conditions in the detector set to give an output about equal to full scale of the V. T. meter take a reading across the smaller part; the ratio of these two readings gives the proper multiplying factor to use with the readings of the meter when connected across the smaller resistance.

EXPERIMENT 46

Object.—Study of the resonance characteristics of the tuned circuit of a radio frequency tuned amplifier, as this is affected by (*a*) the plate resistance of the previous triode and (*b*) the input resistance of the succeeding triode.

Analysis.—One stage of a tuned radio frequency amplifier is shown in Fig. 229. The tuned circuit *L–C* has in itself a certain resistance due to losses in its coil and, to a less extent, in its tuning condenser. With a well-designed coil and a condenser of good manufacture the total resistance should not be more than 1 per cent of the coil reactance; it may be less.

Such a relation of resistance to reactance should give much greater selectivity than is obtained in the actual amplifier, and it is the purpose of this test to find the reason for this discrepancy.

The first effect to be considered is that introduced by the plate circuit of the preceding triode. Referring to Fig. 229, we note that for broadcast frequencies, customary values of the circuit elements give a coil *L* of about 250 microhenrys, a primary coil L_1 of about 10 microhenrys, a coupling between the two of about 60 per cent, and a condenser *C* having a maximum capacity of about 0.0003 microfarad.

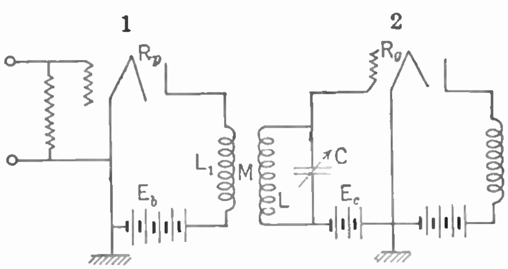


FIG. 229.

A reasonable value for R_p is 5000 ohms, and for R_g we will take 500,000 ohms. A reasonable value for the resistance of the *L–C* circuit at its lowest frequency is 10 ohms and at its highest frequency about 30 ohms.

A coupling of 60 per cent between L_1 and *L* gives for *M* a value of 30 microhenrys. Such a value, with a value of R_p equal to 5000 ohms, gives, for the value of resistance introduced into the *L–C* circuit, about 2.5 ohms at the lower frequency for which the circuit may be tuned and about 20 ohms at the higher frequency.

The value of mutual induction assumed in this problem is not as high as it should be to get maximum amplification; greater coupling between L_1 and L would increase the amplification but also would introduce an objectionably large resistance into the L–C tuned circuit.

The reactance of the condenser C is about 900 ohms for the lower frequency and 2700 ohms for the higher frequency limit of the circuit. The resistance introduced into the tuned circuit by the input circuit is then about 2 ohms at the lower frequency and 15 ohms at the higher frequency.

The resistance of the tuned input circuit then we can tabulate as follows:

Frequency, kilo-cycles	Resistance in circuit itself, ohms	Resistance due to R_p, ohms	Resistance due to R_g, ohms	Total effective resistance of the L-C circuit, ohms
500	10	2.5	1.5	14
1500	30	20	15	65

From these values (which are not at all unreasonable), the width of the resonance curve at the lower frequency limit is 50 per cent wider than the L–C circuit itself would give, and at the upper frequency limit the resonance curve is more than twice as wide as the L–C circuit by itself would give.

To study these effects a circuit is set up as shown in Fig. 229 with the addition of a vacuum tube voltmeter across the condenser C (that is, in parallel with the input circuit of the second triode) and a source of radio frequency power for exciting the L–C circuit. This latter may well be the standard signal generator; a coil of a few microhenrys is connected across the output circuit of the generator, and magnetically coupled to coil L, adjacent to its grounded end. This exciting coil must be coupled to coil L as loosely as feasible, otherwise it too will act to increase appreciably the effective resistance of the L–C circuit.

The vacuum tube voltmeter, the readings of which are to be used in plotting the resonance curve of the L–C circuit under various conditions, will itself introduce an appreciable resistance into the L–C circuit, but this cannot be avoided. Some measuring instrument must be used in getting measurements for the resonance curve, and any such meter must use power and so act to increase the width of the resonance curve of the tuned circuit. The vacuum tube voltmeter consumes less power than any other available instrument.

The L–C circuit is set for its lowest frequency adjustment, and the filaments of both triodes (of Fig. 229) are left cold (no current). The power supply is adjusted for resonance with the L–C circuit, as shown by a maximum indication on the V. T. voltmeter. The coupling between coil L and the power supply coil is then adjusted to give a reading well up on the scale of the V. T. meter.

The frequency of the power supply is then varied over a frequency band sufficiently wide to cover the resonance curve of the L–C circuit. About ten readings of the V. T. voltmeter should be taken at such frequencies that the resonance curve can be accurately drawn.

Another curve is then obtained, in similar manner; with normal current in the filament of the No. 1 triode (Fig. 229). Triode No. 2 still has a cold filament. Another curve is then obtained with the second triode carrying normal filament current.

Three more curves, under the same conditions as above, are then obtained with the L–C circuit adjusted for its high-frequency limit.

If time permits, study the effect with various values of plate voltage on triode No. 1 and various values of grid bias on triode No. 2.

EXPERIMENT 47

Object.—To measure the resistance of a vacuum tube voltmeter at radio frequency.

Analysis.—In many radio tests (such as Exp. 46) a vacuum tube voltmeter is connected across a coil or condenser to measure the voltage drop, and of course as the V. T. meter itself uses some power it disturbs the potential distribution in the circuit to which it is connected. The higher the resistance of the voltmeter, the less is the disturbance its connection to the circuit causes; and also the lower the impedance of that part of the circuit across which the V. T. meter is connected, the less is its effect on the circuit when connected in the manner shown in Fig. 230. The voltmeter, having a resistance R_v, introduces into the L–C circuit an equivalent series resistance equal to X_c^2/R_v. Thus suppose that the true resistance of the L–C circuit is 10 ohms, that the reactance of condenser C is 1000 ohms, and that the vacuum tube voltmeter V has 500,000 ohms resistance. The volt meter will introduce into the L–C circuit 2 ohms resistance, so that the circuit acts as though it had 12 ohms resistance, instead of 10 ohms.

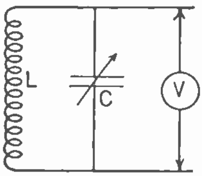


FIG. 230.

By measuring the resistance of the L–C circuit without the voltmeter V connected, and then with it connected, and knowing the reactance of the condenser, the actual resistance of the voltmeter can be calculated.

In spite of the apparent simplicity of this method a closer analysis shows that it is not easy to measure the resistance of the voltmeter with any degree of precision, unless another vacuum tube voltmeter is available; in this case the measurement is easily made, as in Fig. 231.

The L–C circuit is excited from a variable frequency power supply, one of the V. T. voltmeters V being connected across the condenser C. The resonance curve of the L–C circuit (plus voltmeter V_1) is obtained by reading V_1 as the impressed frequency is varied in small steps. Of course the readings of V_1 are not strictly proportional to the current in the L–C circuit, as they should be if a true resonance curve is to be plotted. A resonance curve assumes a constant e.m.f. on the circuit, and ordinates proportional to the current in the circuit. Now as the frequency of the power supply is altered during the test, the voltage induced in the resonant circuit is not constant throughout the run unless the current in the power supply circuit is correspondingly changed. Furthermore, the voltage drop across the condenser C is not quite proportional to the current in the resonant circuit, because its reactance changes with frequency. However, the changes in frequency required to plot a resonance curve are generally so small that both of these errors may ordinarily be neglected.

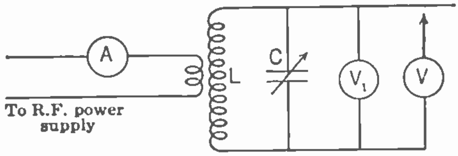


FIG. 231.

Hence, having obtained the resonance curve for the circuit of Fig. 231 (voltmeter V not connected), its resistance is determined from the formula proved in Exp. 6, namely,

$$\delta = \pi \frac{f_2 - f_1}{f_r} = \frac{R}{2f_r L}. \qquad (84)$$

in which f_r is the resonant frequency, f_2 and f_1 are the two frequencies, one on either side of f_r, at which the current drops to $1/\sqrt{2}$ of its resonance value, R is the resistance of the circuit, and L is the inductance.

The voltmeter under test is now connected in parallel to the one already there, and the resonance curve again obtained. From this the new value of circuit resistance is obtained. The excess of this resistance over the value obtained with only one voltmeter connected is the "equivalent series resistance" of the voltmeter under test; let it be called R'.

Now from the resonant frequency f_r and the inductance L, the X of coil (and hence also of the condenser) is calculated. The voltmeter resistance R_v is obtained from the relation

$$R_v = \frac{X^2}{R'}$$

In case only one V. T. voltmeter is available, the resonance curve of the circuit is obtained from the reading of a suitable thermocouple ammeter in the circuit, as in Fig. 232. In this case, however, it is necessary to use a very sensitive millivoltmeter to read the voltage of the thermocouple, as will now be shown.

The ordinary V. T. voltmeter reads only up to about 3 volts. Now if the condenser has 1500 ohms reactance, as we have previously assumed, it can have only 0.002 ampere flowing through it, if the *IX* drop is limited to 3 volts. To get full scale deflection with a current of 0.002 ampere requires a couple with a heater of several hundred ohms, unless some very special galvanometer is used to measure the voltage of the couple. The resonant circuit, without the V. T. voltmeter connected, would thus have 500 ohms resistance, and this will be increased only a few ohms when the voltmeter is connected in parallel. Evidently such a circuit arrangement is entirely unsuitable for this test.

If a 10-ohm couple is used, and a millivoltmeter requiring only 0.25 millivolt for full scale reading (a special portable type of suspension instrument will do this) then full scale reading can be obtained with a current of about 0.008 ampere. Hence 0.002 ampere will give about five divisions of scale, so that a resonance curve will be obtained with reasonable accuracy. The 10-ohm couple together with 10 ohms in the circuit itself gives a total resistance of 20 ohms.

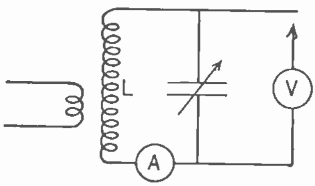


Fig. 232.

Now a 500,000-ohm V. T. voltmeter, across a 1500-ohm reactance, gives an equivalent series resistance of 4.5 ohms, so that the resonance curve would be widened about 20 per cent; such a change could be easily measured, even though not very accurately.

As another expedient the cathode ray oscillograph can be used as a voltmeter to get points for the resonance curve. The 3 volts assumed across the condenser has a maximum value of 4.2 volts, and this voltage on the deflecting plates of the oscillograph will give a considerable deflection, permitting its measurement with reasonable accuracy.

The oscillograph introduces practically no resistance in the resonant circuit, so the width of the resonance curve is determined by the circuit resistance of 10 ohms. When the V. T. voltmeter is connected the resonance curve is widened about 50 per cent, on account of its equivalent series resistance of about 5 ohms.

Measure the resistance of one of the laboratory vacuum tube voltmeters by each of the methods outlined above, and give a brief discussion of their relative results.

If time permits, carry out suitable tests to see if the resistance of the V. T. voltmeter varies with frequency. Make measurements at several radio frequencies, and at two or three audio frequencies. The latter are best made in an a.c. capacity bridge, such as was used in

Exp. 11. The equivalent series resistance of a condenser is measured with and without the V. T. voltmeter in parallel. The difference, by use of the above equation, gives the resistance of the voltmeter. In the bridge test it is possible to determine whether or not the resistance of the voltmeter varies with the amount of voltage impressed upon it. The voltage impressed on the bridge is increased in suitable steps to give voltage on the voltmeter, as measured by itself, at several properly spaced points on its scale.

EXPERIMENT 48

Object.—Measurement of the permeability of iron at various frequencies.

Analysis.—From what experimental results there are available it seems that the actual permeability of iron is independent of frequency, at least up to 10^6 cycles. The apparent permeability, however, falls off very rapidly with increasing frequency, being but little more than unity at radio frequencies, when the iron is made of laminations of ordinary thickness.

The apparent permeability is determined from the flux per ampere turn in the exciting winding, whereas the true permeability is determined from the ratio of flux to *net ampere turns*, the eddy currents in the iron itself being considered in obtaining this net magnetomotive force. At high frequency the eddy currents in the iron core practically neutralize the ampere turns of the exciting winding, thus giving a very small *net magnetizing force* acting on the iron.

It was shown in Exp. 38 that the currents induced in a closed circuit, coupled to one carrying alternating current, increase in magnitude as the frequency is increased, and approach more nearly a phase 180° behind the inducing current.

Now in the case of an iron core coil these secondary induced currents are right in the iron itself; at the higher frequencies they are confined to paths close to the periphery of the iron. The effect of these currents is to give, in the inner part of the iron core, a net magnetomotive force of practically zero value. Hence in the inner part of the core there is no flux, at the higher frequencies.

The only appreciable flux in the iron at the higher frequencies is in a thin layer of iron close to the surface, and this layer becomes ever thinner as the frequency increases.

The action outlined above shows that the total flux in the iron decreases with increasing frequency, even though the ampere turns in the magnetizing coil remain constant. When we define the permeability in terms of " flux per ampere turn " in the exciting winding it follows that this permeability falls off as the frequency increases.

A small iron core coil having an inductance of about 5 millihenrys is suitable for this test. An a.c. bridge with unity ratio is advisable;

this means that the variable standard inductance should have a maximum value of about 5 millihenrys. For frequencies in the audible range, a telephone receiver serves well to indicate a balance, but when higher frequencies are employed, either a sensitive thermocouple, or a heterodyne vacuum tube circuit must be used to indicate balance.

The heterodyne method is indicated in Fig. 233. The grid of the triode is excited by two voltages, that due to the unbalance of the bridge, and a voltage drop across the resistance R_4, which is supplied with power from an oscillating triode. As the frequency impressed on the bridge is altered that supplying the drop across R_4 is also changed, to keep the difference of the two frequencies equal to about 800 cycles per second. To get the most sensitive condition in the detecting triode it is necessary to control the magnitude of the heterodyning frequency voltage; this is done by varying the resistance R_4.

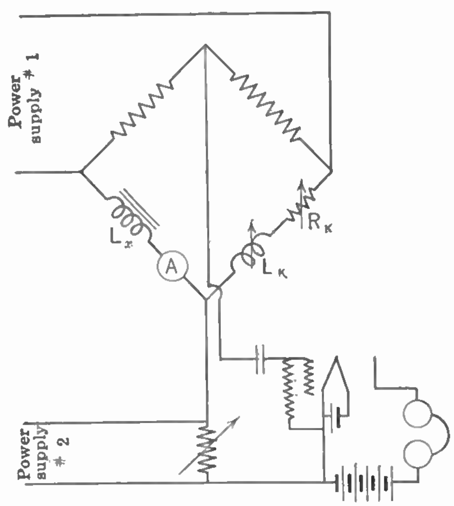


FIG. 233.

A reasonably good balance may be obtained by using a suitable thermocouple. Thus a 150-ohm heater will give full scale reading on the ordinary millivoltmeter with a voltage of 0.75 volt on the heater. By using a more sensitive millivoltmeter (0.24 millivolt full scale, resistance 10 ohms) full scale deflection is obtained with 0.25 volt on the heater. A readable deflection on this millivoltmeter is obtained with a voltage across the heater of 0.025 volt and this will permit a reasonably accurate balance. This scheme, although not as sensitive as the heterodyne triode, is more simple and easier to manipulate.

As the permeability of iron varies with the current in the exciting coil, this should be held constant; this is the purpose of the ammeter A in the circuit of Fig. 233.

With a coil of 5-millihenry inductance, at 1000 cycles the reactance is 30 ohms. A reasonable current to hold on ammeter A is 0.005 ampere; the couple should have a heater of as low a resistance as possible, say not more than 10 ohms.

Up to about 5000 cycles a telephone may be used to balance the bridge; the thermocouple for balancing will not be needed until the signal becomes inaudible. At the lower frequencies the voltage impressed on the bridge will be low, so that the thermocouple balance would be very insensitive; here, however, where the thermocouple is insensitive as a balance indicator, the telephone receiver is used.

At 1000 cycles a 5-milliampere current through the coil requires a voltage on the bridge of about 0.3 volt. At 5000 cycles this must be increased to 1.5 volts, if the inductance is still 5 millihenrys. With this voltage on the bridge the thermocouple balance will be rather poor. However, as the frequency goes up, more voltage is required on the bridge, so the balance becomes more sensitive. Thus at 50 kc. about 7.5 volts are required, and at 100 kc. about 15 volts. These figures are based on the inductance of the coil remaining constant; in so far as the inductance of the coil drops, a smaller voltage will be required on the bridge.

After a preliminary run it may be found that a current of 0.01 or 0.02 ampere is more suitable than the 0.005 ampere suggested. At the highest frequencies the voltage impressed on the bridge should be as high as the power source will conveniently give.

Still another way of measuring the inductance is to set up a resonant circuit. Having a variable condenser, of very wide range, in series with the coil and a thermocouple ammeter, the capacity is varied, at any desired frequency, until resonance is obtained. When the frequency and capacity are known, the inductance is at once calculated. When this method is used it is well to have a short air gap in the magnetic circuit, to make the resonance frequency more definitely fixed.

The " constant impedance " method, as outlined in the latter part of Exp. 36, may also be used. In this method the same current is obtained through the coil, with and without a variable series condenser, and from the known value of the series condenser and frequency the reactance of the coil is calculable.

Having measured the reactance, or inductance, of the coil by any one of the methods outlined, the permeability is calculated from the known number of turns in the coil and dimensions of the iron core.

Using the resonance method, bridge method, and constant impedance method, measure the inductance of an iron core coil for frequencies extending from the audio range to 100 kc. or more. Try one coil with a core of thin laminae, and another coil with an " iron dust " core; this is the material used in telephone loading coils.

EXPERIMENT 49

Object.—Study of modulation schemes. Effect of modulation upon frequency of carrier wave. Measurement of side band width, to check with modulation frequencies.

Analysis.—One of the earliest modulation schemes to be used operated on the idea of varying the resistance in the oscillating circuit; the resistance was modulated at voice frequency by using a microphone as the variable resistance. The idea is illustrated in Fig. 234; it is still used by some of the amateur radio telephone transmitters.

The triode oscillates at a frequency fixed by the L_1–C_1 circuit, the grid being excited by the coil L_2. The coil L_3 in the antenna circuit is variable, and serves for tuning the antenna circuit to the triode frequency. The microphone M is directly in series with the antenna, so that the current in the antenna will increase and decrease as the voice affects the microphone resistance. This scheme has three disadvantages: it is inefficient from the power standpoint, it can handle only small amounts of power, and it sends out a variable frequency. The variation in microphone resistance will affect to some extent the frequency of the current in the L_1–C_1 circuit, that is, the carrier current of the transmitter. Although hardly noticeable with a low-frequency carrier, it becomes very objectionable when the carrier frequency is measured in megacycles.

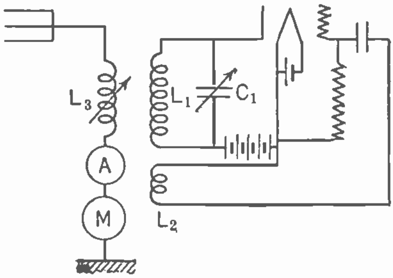


FIG. 234.

This effect can be studied by setting up a circuit similar to that of Fig. 234, using a dummy antenna tightly coupled to the L_1–C_1 circuit, and a variable resistance in place of the microphone. This resistance should have a maximum value sufficient to reduce the current in the antenna ammeter to nearly zero.

Attempt to measure the change in frequency by wave meter; it may be so small that the method fails. In that case set up an oscillating receiving circuit close enough to the transmitter circuit to produce a beat note. Adjust the receiver frequency to make this beat note go to zero. Now vary the antenna resistance and get the change in frequency of the carrier by the pitch of the beat note which will be heard. Get the relation between frequency change and antenna circuit resistance. Have the coupling between transmitter and receiver sufficiently loose that the synchronizing action spoken of in Exp. 31 is not noticeable.

Modulation in commercial radio phone transmitters is accomplished by impressing the modulating voltage on the grid or plate of the oscillating tube which is generating the carrier frequency. When the modulating voltage is impressed upon the plate of the oscillator, the scheme is called plate modulation; this is the more common scheme and is the one which will be studied in this test.

To measure, quantitatively, the energy distribution in a modulated wave is not simple, because of limitations in apparatus as ordinarily available. A modulated wave has most of its power in the carrier; a *little* power spreads out into the radio spectrum from both sides of the carrier, the amount of this power depending upon the amount of modulation, and the degree of spread depending upon the frequency of the modulation frequency.

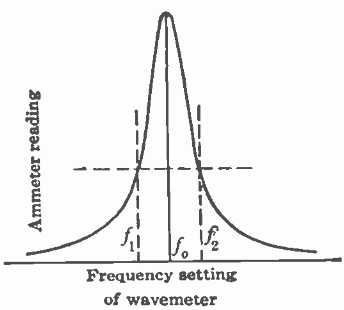


Fig. 235.

We shall use a wave meter to analyze the complicated wave forms of the modulated transmitter, so the first question to be investigated is the performance of the wave meter itself, for a simple, unmodulated carrier wave. In Fig. 235 such a response curve is shown. The carrier frequency is f_0, and as the tuning of the wave meter is changed, the response of its indicator follows the curve shown. The ordinary thermocouple ammeter used in wave meters is equipped with a uniform scale millivoltmeter, so that its readings are proportional to the square of the current, hence the ordinates in Fig. 235 are proportional to $(\text{current})^2$.

Now, as shown in Exp. 6, a series resonant circuit (this is all there is to a wave meter) shows a form of resonance curve which depends upon the resistance; quantitatively the two frequencies f_2 and f_1 (of Fig. 235) at which the (current) is half of the $(\text{resonance current})^2$ are separated from each other by an amount determined from the resistance according to the relation

$$\delta = \frac{R}{2f_rL} = \pi\frac{f_2 - f_1}{f_r} \quad \text{or} \quad f_2 - f_1 = \frac{\delta}{\pi}f_r. \quad . \quad . \quad . \quad (85)$$

Now a good wave meter may have a decrement as low as 0.015, so that we may assume for a good wave meter $f_2 - f_1 = 0.005 f_r$.

Thus if the frequency f_r was 100 kc., the two frequencies f_2 and f_1 would be separated by 500 cycles; f_2 would be 100,250 cycles and f_1 would be 99,750 cycles.

Now suppose that the carrier wave was modulated 100 per cent at 250 cycles. The wave meter response would be controlled by the carrier current of 100 kc. and two side band currents, each half the magnitude of the carrier. But the wave meter response is proportional to the square of the current, so the actual response curve for this modulated current would be made up as shown in Fig. 236. The response for the two side bands, each one-quarter as large as the response for the carrier, would be hidden in the carrier resonance curve, and the response would not be perceptibly different from that of an unmodulated carrier.

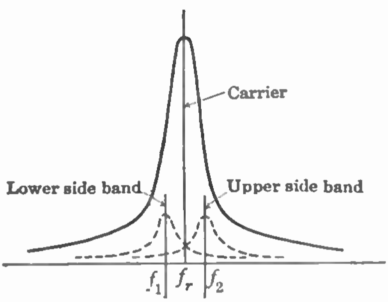


FIG. 236.

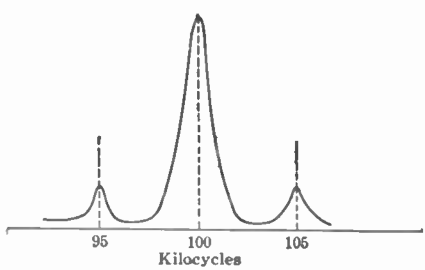


FIG. 237.

However, suppose that the modulation frequency is 5000 cycles; one side band will be at 105 kc. and the other at 95 kc., and the wave meter response would be as shown in Fig. 237. Here the two side bands are distinctly shown.

A little study of the above analysis shows that for a given frequency of modulation, the side band resonance curves stand out from the carrier resonance curve more distinctly as the frequency of the carrier is lowered. Thus the $f_2 - f_1$ width of the carrier resonance curve, for a wave meter with $\delta = 0.015$, is about 500 cycles when the carrier is 100 kc., but the width is only 125 cycles when the carrier frequency is 25 kc.

Hence in order to show up the modulation frequency action as distinctly as possible it is advisable to use a very low carrier frequency, and this test is first to be carried out at 25 kc. The first scheme to be investigated is that using plate modulation, frequently called Heising modulation. In this scheme the voltage on the plate of the oscillator is made to increase and decrease about its normal value, at the frequency of modulation. For complete modulation, or 100 per cent modulation as it is called, the amplitude of the modulation voltage must be equal to the normal unmodulated plate voltage of the oscillator.

In the Heising scheme of modulation the oscillator triode draws its plate current through an iron core inductance choke, of several henrys inductance. Another triode, of the same (or somewhat larger) rating as the oscillator, also draws its plate current through the same choke; the second tube is called the modulator.

The complete lay-out of apparatus suitable for this test is given in Fig. 238, in which tube *C* is the oscillator and tube *B* is the modulator. It will be seen that they both draw plate current through the same choke coil.

The oscillator *C* has a tuned circuit, L_1–C_1, in the plate, and the grid is excited by the grid coil L_2. A grid condenser *C* and leak resistance

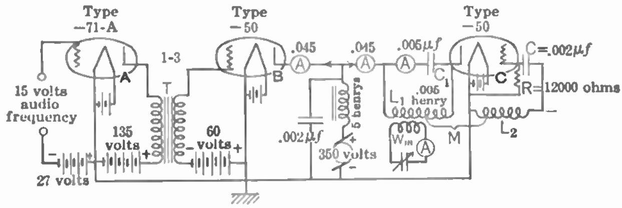


FIG. 238.

R serve to control the amplitude of oscillations. The wave meter *W* is coupled to coil L_1; its readings show the composition of the current which the L_1–C_1 circuit would supply to an antenna if one were used.

The modulator tube *B* is of the same rating as the oscillator tube; in Fig. 238 the voltage and current indicated assume that both modulator and oscillator tubes are type −50 triodes. The grid of the modulator is supplied from the secondary of a 1–3 transformer in the plate circuit of a type −71-A triode. The grid circuit of this tube is shown in the diagram to be connected to a source of variable-frequency a.c. power, having a voltage amplitude up to 15 volts. This amount of voltage is available from the output circuit of the audio frequency oscillator with which the average radio laboratory is equipped. The frequency available from the ordinary oscillator ranges from 100 to 100,000 cycles, but in this test we shall not require frequencies above about 5000 cycles.

The oscillator circuit L_1–C_1 is shown made up of 5 millihenrys inductance and 0.005 microfarad of capacity. It is well if this condenser is of mica; paraffin paper is too likely to break down in a test of this kind.

If the audio frequency oscillator supplied 15 volts to the input of the voice frequency amplifier (−71-A triode), the input to the grid of the modulator tube will be $15 \times 3 \times 3 = 135$ volts. As the amplification factor of the modulator triode is 3.8, this voltage is equivalent to $135 \times 3.8 = 513$ volts acting in the plate circuit of the modulator. This is much more than is required to give 100 per cent modulation; hence it will not be necessary to use as much as 15 volts input to the −71-A tube.

In the first test the audio frequency input is to be set at 5000 cycles, and adjusted in amplitude until the A. F. a.c. voltage across the choke coil L is equal to $1/\sqrt{2}$ of the voltage used for the plate supply of tubes B and C. This gives about 100 per cent modulation, the plate current of the oscillator (average value) varying from zero to 0.09 ampere at a frequency of 5000 cycles. The current in each of two side bands will now be half as great as the current in the carrier frequency component. The wave meter is coupled closely enough to the L_1–C_1 circuit, so that when tuned to the carrier frequency its meter reads full scale. Its response should then have the form given in Fig. 239; the two side bands should show up as distinct resonance curves.

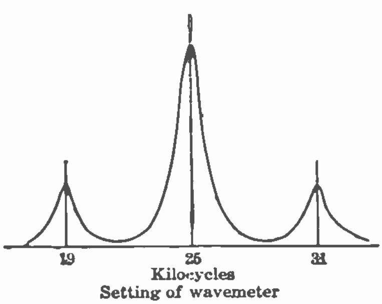


FIG. 239.

Keeping the amplitude of the modulation voltage the same, the frequency is now decreased in suitable steps (and a wave meter response curve obtained for each) until the two side bands merge into the carrier frequency resonance curve and disappear.

Next with the modulation voltage set at some suitable value, such as 5000 cycles, the amplitude of the modulation voltage is changed in several steps and the wave meter response curve obtained for each. Try one or two values of voltage greater than is required to give 100 per cent modulation. Repeat these runs with the carrier frequency changed to 100 kc.

To use grid circuit modulation the modulator tube B is disconnected and the choke coil L is removed from the plate circuit of the oscillator. The secondary of transformer T is inserted in the grid return wire of the oscillator next to the filament. The average voltage of the grid of the oscillator is then forced to follow the modulation frequency voltage, and so of course the average value of the plate current of the oscillator follows this voltage. This results in a radio frequency current in the L_1–C_1 circuit which has an amplitude following the modulation voltage.

With the L_1–C_1 circuit set to generate 25 kc., make two sets of response curves, one for the conditions of variable amplitude modulation voltage and fixed frequency, and the other with fixed amplitude and variable frequency.

EXPERIMENT 50

Object.—Study of superheterodyne detector. Variation of intermediate frequency signal with strength of local signal. Action of intermediate frequency transformer.

Analysis.—In the double detection receiver, or superheterodyne receiver as it is generally called, the high-frequency signal received by the antenna is supplied to the same detector input circuit, in which there exists a constant amplitude, locally generated signal which has a frequency 50 or perhaps 100 kc. different from the frequency of the incoming signal. This action is exactly the same as the ordinary heterodyne receiver shown in Fig. 240, which was studied in Exp. 31, with the exception that the beat frequency is not audible as it is in the ordinary heterodyne receiver, but above audibility, hence the common name, superheterodyne.

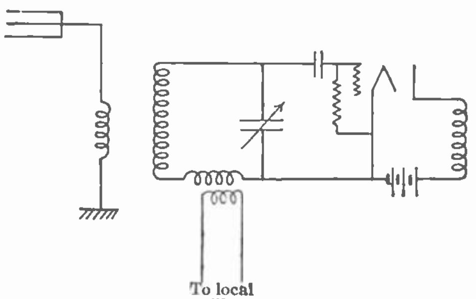


Fig. 240.

In Exp. 31 the local oscillation was produced by the detector triode itself; the input circuit has such L–C product that the natural frequency differed from that of the incoming signal by a few hundred cycles. A feed back coupling from the plate circuit maintained these local oscillations.

Now to receive a radio signal with any reasonable efficiency the input circuit should be tuned to the incoming signal; evidently in a heterodyne receiver this is impossible, because if it were so the local oscillations would have the same frequency as the incoming signal and there would be no beat frequency to give the audible signal. However, even if the input circuit is not tuned to the incoming signal in the heterodyne scheme of reception, it is only a few hundred cycles away from resonance, a difference from the tuned condition so slight that but little loss occurs because of the mistuning.

In the superheterodyne receiver the difference of the two frequencies is, say, 100 kc.; a receiving circuit adjusted for resonance 100 kc. away from the signal frequency would, however, pick up but little energy from the antenna, so that the system would be very insensitive. For this scheme of reception the local oscillations must be generated by another triode, and supplied to the detector input circuit (which is accurately tuned to the signal frequency) by an inductive or capacitive coupling. The arrangement then is as shown in Fig. 241. The input circuit to the detector triode T_1 is made up of L_1, L_2, and C_1, and this circuit is tuned to the signal being received by the antenna, say 1000 kc. The triode T_2, the local oscillator has its grid circuit L_3, L_4, C_2

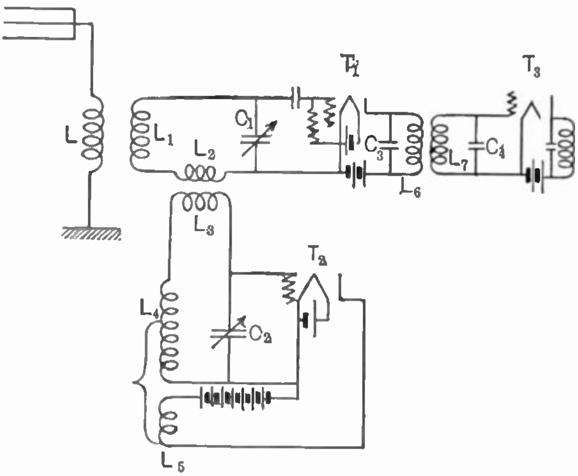


FIG. 241.

tuned for either 1050 or 950 kc.; this circuit is kept in oscillation by the tickler coupling, L_4–L_5. The amount of this local signal which is supplied to the detector input circuit is controlled by the coupling between L_3 and L_2.

Let us say the local oscillator is set to give 950 kc. The beat frequency is 50 kc., and this is rectified by the detector action. A condenser C_3 around the coil L_6 in the plate circuit of T_1 serves to by-pass the high-frequency currents. The 50-kc. current, however, is passed along to the circuit L_7–C_4 which is broadly tuned to 50 kc. This is the first stage of the intermediate frequency amplifier (I. F. A.); in the average superheterodyne there are several of these I. F. A. stages.

As mentioned before, any other intermediate frequency, such as 75 or 100 kc., may be used instead of 50 kc.; in this case the local oscillator is set to generate a frequency either 75 or 100 kc. distant from that of the received signal. Also the I. F. A. transformers would be tuned (broadly) for 75 or 100 kc., as the case might be. And it will also be noted that the frequency of the local oscillations may be either higher or lower than that of the signal.

Now the interesting feature of this type of amplification lies in the fact that any modulation which may be present in the signal wave is passed along on the rectified beat frequency, and so is passed along to the I. F. A. The advantage of this scheme of amplification lies in the fact that it is easier to amplify a 50-kc. current than a 1000-kc. current; the 50-kc. amplifier is much more stable and does not require the special scheme for neutralizing the inter-electrode capacity feed back which is so troublesome in the triode high-frequency amplifier.

Furthermore, the I. F. A. is tuned to the same frequency, whatever may be the frequency of the signal, hence the I. F. A. inter-stage transformer may be built with fixed condensers. The only condensers which must be adjustable are those in the detector input circuit and in the local oscillator circuit, whereas the ordinary type of radio frequency amplifier requires from four to six accurately matched variable condensers.

One of the questions to be investigated in this test has to do with the magnitude of the local oscillation supplied to the detector input; how does the strength of the beat frequency signal, passed on to the I. F. A., vary as the amount of local signal supplied to the detector is varied? Does the proper magnitude of this local signal in the detector circuit depend upon the strength of signal supplied by the antenna? Furthermore, with the type of I. F. A. transformer generally used, how does the action of the I. F. A. depend upon the frequency of the local oscillations, all other factors remaining the same.

Arrange a circuit as shown herewith (Fig. 242); the standard signal generator is to be connected in the detector circuit at A. The input circuit L_1–L_2–C is to be tuned for the signal frequency. A vacuum tube voltmeter is to be connected across the secondary of the I. F. A. transformer.

With a reasonable signal strength, and the local oscillator set to give its proper frequency, take a series of readings between amount of local signal supplied to the detector circuit and the reading of the V. T. voltmeter. The amount of local signal supplied can be measured by using triode T_1 as a voltmeter, the readings of ammeter A_p having been calibrated in terms of signal voltage introduced in the L_1, L_2, C circuit, by a preliminary test. During this calibration of T_1 in terms of the strength of local signal put into the L_1, L_2, C circuit, the input from the standard signal generator is set at zero.

With a reasonable coupling between L_2 and L_3 (as determined in the previous test), get a series of readings between strength of signal, as supplied in known amounts by the standard signal generator, and the reading of the vacuum tube voltmeter.

With reasonable signal strength and reasonable amount of local oscillation supplied to the detector get the relation between the reading of the V. T. voltmeter and the frequency difference of the two signals in the detector circuit. Either the frequency of the standard signal generator may be held constant, and the local oscillator frequency changed, or vice versa.

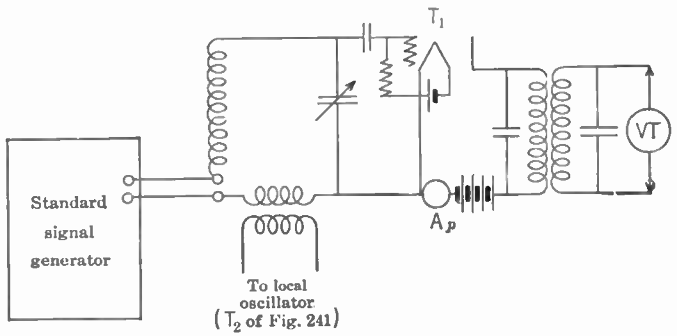


FIG. 242.

With fixed signal strength, find the effect on the reading of the V. T. voltmeter of using various amounts of local signal in the input circuit of triode T_1.

In working with this circuit the student is to notice two defects in this circuit arrangement, which sometimes are very troublesome. If the local oscillator is set for 900 kc. and the input circuit tuned to 950 kc. to pick up a distant station on this frequency, a local station on 850 kc. may entirely drown out the distant station. Also there will be two settings of the local oscillator, 950 kc. and 1050 kc., which are equally good for receiving the 1000-kc. station. By incorporating in the superheterodyne properly tuned circuits, mechanically operated from the main tuning control, both these defects can be overcome.

Of course in a properly designed superheterodyne receiver precautions are taken to prevent the locally generated high frequency power from being sent out from the antenna. Ordinarily a stage of neutralized R.F. amplification (see Ex. 51) is used for this purpose.

EXPERIMENT 51

Object.—A study of neutralization of the effect of inter-electrode capacity in triode circuits.

Analysis.—The use of the screen grid tube has very largely done away with the need of neutralizing the effect of the grid-plate capacity in the ordinary triode. However, as there are still many triodes in use in radio frequency amplifier circuits, and the necessity for neutralizing capacity feed back may exist in many vacuum tube amplifier circuits, the following analysis and tests may be of much service.

In Fig. 243 is shown an ordinary arrangement of split coil oscillating circuit. The frequency is fixed by the *L–C* circuit; the possibility

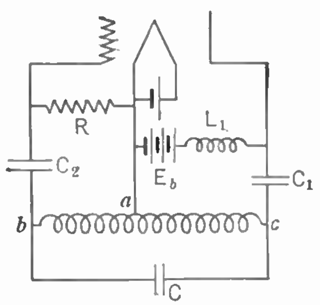


Fig. 243.

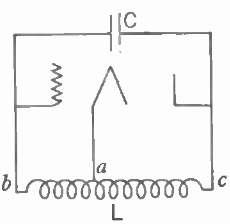


Fig. 244.

of oscillation and amount of oscillatory power produced are largely determined by the relative positions of the coil taps *a*, *b*, and *c*. The high-voltage supply to the plate is by what is often called parallel feed; the plate power supply E_b is in series with a radio frequency choke coil L_1 to keep the radio frequency power out of this circuit. The condenser C_1 (of comparatively large capacity) keeps the high c.c. voltage of the plate supply out of the oscillatory circuit, and the condenser C_2, in cooperation with the leak resistance *R*, acts to limit the power input to the oscillating triode.

Now this circuit, in simplified form, is redrawn in Fig. 244. The power supply circuits have been omitted. The frequency of oscillation is again fixed by the *L–C* circuit. The capacity *C*, it will be noted, is connected directly between the grid and plate of the triode. This circuit will oscillate even if there is no mutual induction between the section *ab* of the coil and section *ac*.

Now there is always an appreciable capacity between the grid and plate of a triode, from 5 to $10\mu\mu f$ in small tubes up to $100\mu\mu f$ or more in power tubes. So that with only two coils, one between filament and grid, and the other between filament and plate, the circuit is electrically equivalent to that of Fig. 245 and will generally oscillate and the frequency fixed by

$$f = \frac{1}{2\pi\sqrt{(L_1 + L_2)C_{gp}}}.$$

Now many triode circuits not intended to oscillate, do have one or more possible frequencies of oscillation; in Fig. 246 such a one is shown.

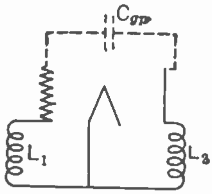


Fig. 245.

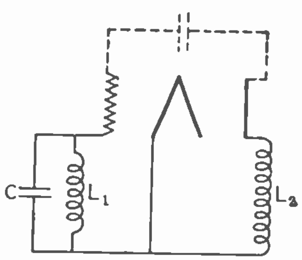


Fig. 246.

This circuit, if it oscillates at all, will have its frequency fixed by the L_1–C circuit. A certain minimum value of L_2 is required to produce oscillations, and L_2 must not be so large that its natural frequency is lower than the frequency of the L_1–C circuit. Furthermore, as the amount of power which can be fed to the L_1–C circuit from the plate circuit, through the condenser C_{gp}, is small, it is necessary that the decrement of the L_1–C circuit be reasonably low if oscillations are to occur.

It will be recognized that the arrangement of Fig. 246 is one stage of the ordinary tuned radio frequency amplifier, and in such an amplifier it is true that the decrement of the L_1–C circuit is low, and that the value of inductance in the plate circuit is plenty to satisfy the requirements for oscillation of the L_1–C circuit, through the capacitive feed back between grid and plate of the triode.

It is ordinarily said that the triode is a " one-way repeater," that oscillations occurring in the input circuit will be passed along to the plate circuit, whereas oscillations in the plate circuit will not be passed back to the grid circuit. From the foregoing simple analysis it is apparent that, owing to the inter-electrode capacity, energy can return from plate to grid, so that the action is not strictly " one-way." To make the triode a truly one-way device the neutralizing circuit to be discussed below must be used.

The fundamental idea involved in all the various circuit arrangements which have been suggested for neutralizing the grid-plate feed back is the same, namely, to arrange an auxiliary circuit in such a manner that another feed back path is provided, having the voltage feed back of this circuit *equal to that through the inter-electrode capacity and opposite in phase.* The net voltage fed from the plate circuit back into the grid circuit, or vice versa, is thus made equal to zero.

In Fig. 247 is shown one form of neutralizing circuit; the filament of the triode is connected to the mid-point of the coil in the tuned input circuit, and the extra condenser, C_1, is connected between the lower end of coil L and the plate.

If then the plate voltage acting through C_{gp} tends to make a current i flow downward through coil L the same change in voltage will, through condenser C_1, tend to make a current i flow upwards in coil L. These two effects, acting simultaneously, just neutralize one another.

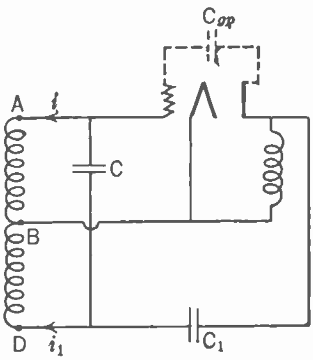


FIG. 247.

If the filament connection is made to coil L at some point other than the mid-tap, then condenser C_1 will have to be either larger or smaller than C_{gp}. The lower the filament connection is made on coil L, the larger must be the condenser C_1.

If N_1 is the number of turns between the filament tap and lower end of the coil and N_2 is the number between the filament tap and upper end of the coil, then the relative capacities of the two condensers are given by the relation

$$\frac{N_1}{N_2} = \frac{C_{gp}}{C_1}. \qquad (86)$$

To make this scheme of neutralization operate well the coil L must be of compact construction, so that the coupling between the two parts of the coil is reasonably tight. For perfect neutralization the coupling between the two coil sections $A\ B$ and $B\ D$ (Fig. 247) should be 100 per cent, an obviously impossible figure. In practice 70 per cent, or even 60 per cent, proves sufficient to make the neutralization satisfactory.

In another scheme, shown in Fig. 248, the extra coil to be used for the neutralization circuit was connected in the plate circuit. The part $B\ D$ of the plate coil connects to the grid through the neutralizing condenser C_1. (The neutralizing condenser is made adjustable in whatever part of the circuit it is used.) For this connection also it is not necessary that $C_{gp} = C_1$, but the relation must be satisfied that

$$\frac{N_1}{N_2} = \frac{C_{gp}}{C_1}.$$

In this arrangement the magnetic coupling between coils $A-B$ and $B-D$ must be as tight as feasible; in some of the better sets the two coils have been arranged to have 80 per cent coupling.

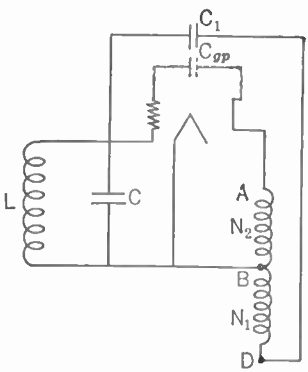


Fig. 248.

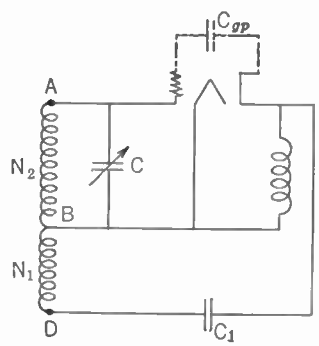


Fig. 249.

In still other circuit arrangements the coil to which the neutralizing condenser was connected was in the input circuit of the next triode of the amplifier; the only requirement is that the magnitude and phase of the voltage at the point the neutralizing condenser is connected are correct.

The scheme shown in Fig. 247 has not been used very much because it neutralizes no better than the other schemes, and it cuts down very materially the signal impressed upon the triode input circuit; with the mid-tap as shown only half of the voltage across the tuning condenser is impressed on the grid of the triode. An equally efficacious arrangement of utilizing the input circuit for giving the proper neutralizing voltage is shown in Fig. 249; in this scheme all of the signal voltage is impressed on the triode input circuit. Here also the coupling between N_1 and N_2 should be as high as possible.

In trying this experiment there must be no magnetic coupling between the input and output circuits of the triode; such coupling cannot be properly neutralized by a capacitive coupling, and of course vice versa.

To carry out the test easily, the neutralizing condenser C_1 should be a very small vernier condenser with a long insulating handle. The coils should be arranged at right angles to each other, as shown in Fig. 250, to reduce the mutual induction to zero. Plug-in coils, with suitable receptacles mounted on the test board, are best. Several coils with different ratio of turns between the two sections should be tried, the neutralizing coil being first in the grid circuit and then in the plate circuit. A milliammeter in the plate circuit serves to indicate oscillations; when the neutralizing condenser has been adjusted to the proper value oscillations cease and the reading of the milliameter increases. If the neutralizing condenser is either too large or too small, oscillations will occur; the proper value of neutralizing condenser is mid-way between the two values, high and low, for which oscillations cease.

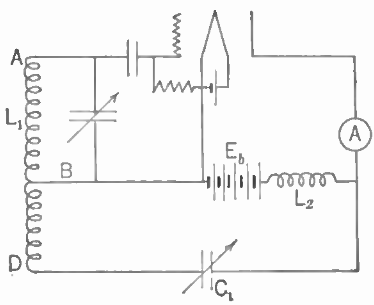


FIG. 250.

After proving the ideas outlined in the analysis, try the effect of having a weak coupling between the sections *A–B* and *B–D*. It will probably be found that very high-frequency oscillations occur, in part of the circuit only, when the voltage E_b is high. Their frequency can of course be measured by wave meter. It will be found possible to eliminate them by putting a resistance in series with the grid connection (between grid and input coil). Generally 100 ohms will be sufficient to eliminate these so-called " parasitic " oscillations.

In carrying out this test it is well to select the L and C to give a natural frequency of 1000 kc. or more. The tendency to oscillate decreases as the capacity in the tuned circuit is increased; it may be found that with a capacity of say $0.0003\mu\mu f$ in the tuned circuit no oscillations occur even if the neutralizing winding is not connected.

It is with small values of tuning condenser ($0.0001\mu\mu f$ and less), and high values of E_b voltage, that the inter-electrode capacity causes most trouble.

LIST OF EXPERIMENTS

(1) Measurement of self-induction and resistance at power frequencies. Iron core coil and air core coil.

(2) Measurement of mutual induction at power frequencies. Dependence upon coil separation and position. Coefficient of coupling.

(3) Measurement of capacity at power frequencies. Variation with frequency and voltage. Power loss. Effect of voltage wave form. Paper condensers and electrolytic condensers.

(4) Measurement of self-induction and resistance of a coil, to which another coil is coupled. Effect of resistance of second circuit, frequency of current, and coupling.

(5) Measurement of reactance and resistance of a circuit to which is coupled another coil in series with a condenser. Effect of varying frequency, coefficient of coupling, and resistance of second circuit.

(6) Resonance in a circuit containing coil and condenser in series. Effect of resistance on shape of resonance curve. Calculation of decrement from shape of resonance curve, and from the constants of the circuit. Effect of putting voltmeter across the condenser.

(7) Study of parallel resonance. Effect of resistance upon shape of resonance curve. Effect of harmonics in voltage wave. Effect of supplying power to a tapped portion of the coil.

(8) Resonance relations in a circuit similar to a tuned radio frequency amplifying circuit. Effect of resistance shunted across the condenser.

(9) Resonance in magnetically coupled tuned circuits. Effect of coupling. Effect of mis-tuning.

(10) Measurement of self-induction, resistance, and mutual induction by bridge method, at audio frequency; by comparison with variable standard inductance or capacity. Measurement of resistance of standard inductance, at various settings and frequencies, by resonance bridge.

(11) Use of a.c. bridge to measure capacity and phase angle at audio frequencies. Measurement of specific inductive capacity and phase angle of typical insulating materials.

(12) Measurement of internal capacities of a vacuum tube, by special bridge.

(13) Measurement of characteristics of a horn-type loud speaker by a.c. bridge. Effect of clamping the diafram. Effect of horn. Dynamic speaker with and without exciting field.

(14) Calibration of hot wire meter, and thermocouple meter on c.c. circuit. Variation of heater resistance with current. Efficiency of conversion in the thermocouple instrument.

(15) Study of rectifiers, by continuous current, and alternating current characteristics. Use of crystal rectifier, gas rectifier, and copper-oxide rectifier.

(16) Use of buzzer wave generator. Construction of various types of wave meters. Various resonance indicators. Measurement of inductance and capacity by wave meter.

(17) Measurement of antenna characteristics by buzzer excited wave meter and aperiodic detecting circuit. Relation between antenna capacity and number of wires. Effect of loading coils, and series condenser, upon wave length.

(18) Operation of a simple spark transmitter. Effect of antenna resistance and coupling upon energy distribution of radiated power.

(19) Study of electron emission from various hot surfaces. Efficiency of different emitters.

(20) Study of the c.c. characteristics of various types of triodes.

(21) Study of c.c. characteristics of tetrodes and pentodes.

(22) Use of the triode as a voltmeter. With and without a grid condenser. Effect of frequency upon accuracy of calibration. Use of compensating circuit to reduce "zero signal" reading of plate circuit ammeter to zero.

(23) Use of triode as detector with various circuit arrangements. Comparison with crystal rectifier.

(24) Measurement of plate circuit resistance and amplification factor of a triode under various conditions.

(25) Measurement of capacity and resistance of the input circuit of a triode, under various conditions.

(26) The triode as a power converter, using separate excitation. Tuned and untuned output circuits.

(27) The triode as a self-excited converter in a typical oscillating circuit.

(28) Measurement of resistance and reactance of a tuned circuit as these are affected by the regenerative action of a triode. Bridge measurement.

(29) Measurement of coil resistance at radio frequency. Effect of frequency, wire size, and wire spacing.

(30) Measurement of inductance, capacity, and resistance of an

antenna, at various frequencies, using oscillating triode as power supply.

(31) Action of a triode in a typical receiving circuit arranged for heterodyne reception. Tendency to synchronism with another oscillator.

(32) Quantitative study of conditions for setting up oscillations in four typical oscillating circuits. Comparison with conditions predicted from theory.

(33) To measure the amplification of a single stage of audio frequency amplification, using resistance, inductance, and transformer repeating circuits. Effect of resistance and capacity across the secondary of the transformer. Effect of reversing connection of secondary winding.

(34) Calibration of a wave meter from a standard meter. Check against a circuit made up of standard L and C. Check against quartz oscillators. Check against Bureau of Standards signals.

(35) To measure the decrement of a wave meter by variation of the impressed frequency, and by variation of the wave meter capacity.

(36) To measure the effective a.c. permeability of iron at audio frequencies, and how this varies with magnitude of alternating current, frequency, and superimposed continuous current magnetization.

(37) To study the effect of dead ends of a tapped coil. Effect of short circuiting the unused portion.

(38) Measurement of shielding of one circuit by another. Effect of resistance in the shielding circuit. Effect of frequency. Shielding by metal plates, of various materials and thicknesses.

(39) Study of cathode ray oscillograph and its use in measuring transformer ratios, hystersis, and frequency ratios.

(40) Measurement of amplification, selectivity, and fidelity of a complete broadcast receiver.

(41) Study and calibration of a c.c. amplifier.

(42) Measurement of the internal capacity of coils of various types. Relation of capacity to size of coil and kind of winding. Natural frequency of a coil.

(43) Study of filters of various types. Effect of resonant shunts on sharpness of cut off. Effect of using several sections.

(44) Attenuation characteristics of a radio frequency band pass filter.

(45) Study of detector action. Relation of A. F. voltage output to R. F. input, with constant modulation. Effect of varying degree of modulation and frequency of modulation.

(46) Action of the tuned circuit of a R. F. amplifier, as this is affected

by the plate circuit of the preceding triode and by the input circuit of the succeeding triode.

(47) Measurement of the resistance of a vacuum tube voltmeter.

(48) Variation of the apparent permeability of iron as the frequency is raised to high values.

(49) Study of modulation. Resistance modulation. Plate circuit modulation. Grid circuit modulation. Energy distribution in a modulated wave.

(50) Study of superheterodyne detector. Variation of intermediate frequency signal with strength of local signal. Action of intermediate frequency transformer.

(51) A study of neutralization of inter-electrode capacity coupling in a triode amplifier.