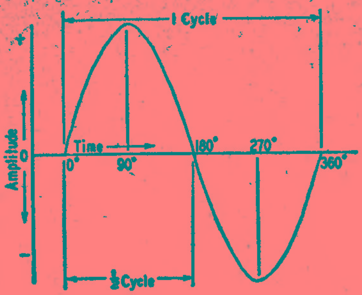


Fig. 2-23 — An ac cycle is divided off into 360 degrees that are used as a measure of time or phase.

Measuring Phase

The phase difference between two currents of the same frequency is the time or angle difference between corresponding parts of cycles of the two currents. This is shown in Fig. 2-24. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees earlier in time. It is equally correct to say that *B lags A* by 45 degrees.

Two important special cases are shown in Fig. 2-25. In the upper drawing *B* lags 90 degrees behind *A*; that is, its cycle begins just one-quarter cycle later than that of *A*. When one wave is passing through zero, the other is just at its maximum point.

In the lower drawing *A* and *B* are 180 degrees out of phase. In this case it does not matter which one is considered to lead or lag. *B* is always positive while *A* is negative, and vice versa. The two waves are thus *completely* out of phase.

The waves shown in Figs. 2-24 and 2-25 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage and *B* current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the total or resultant current (or voltage) also is a sine wave, because adding any number of sine waves of the same frequency always gives a sine wave also of the same frequency.

Phase in Resistive Circuits

When an alternating voltage is applied to a resistance, the current flows exactly in step with the voltage. In other words, the voltage and current are in phase. This is true at any frequency if the resistance is "pure" — that is, is free from the reactive effects discussed in the next section. Practically, it is often difficult to obtain a purely resistive circuit at radio frequencies, because the reactive effects become more pronounced as the frequency is increased.

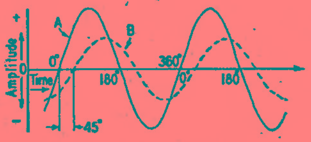


Fig. 2-24 — When two waves of the same frequency start their cycles at slightly different timas, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) latar than wave *A*, and so lags 45 degrees behind *A*.

In a purely resistive circuit, or for purely resistive parts of circuits, Ohm's Law is just as valid for ac of any frequency as it is for dc.

REACTANCE

Alternating Current in Capacitance

In Fig. 2-26 a sine-wave ac voltage having a maximum value of 100 volts is applied to a capacitor. In the period *OA*, the applied voltage increases from zero to 38 volts; at the end of this period the capacitor is charged to that voltage. In interval *AB* the voltage increases to 71 volts; that is, 33 volts additional. In this interval a *smaller* quantity of charge has been added than in *OA*, because the voltage rise during interval *AB* is smaller. Consequently the average current during *AB* is smaller than during *OA*. In the third interval, *BC*, the voltage rises from 71 to 92 volts, an increase of 21 volts. This is less than the voltage increase during *AB*, so the quantity of electricity added is less; in other words, the average current during interval *BC* is still smaller. In the fourth interval, *CB*, the voltage increases only 8 volts; the charge added is smaller than in any preceding interval and therefore the current also is smaller.

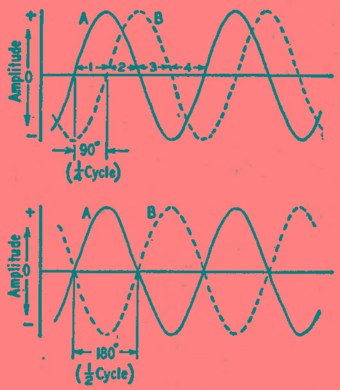


Fig. 2-25 — Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

By dividing the first quarter cycle into a very large number of intervals it could be shown that the current charging the capacitor has the shape of a sine wave, just as the applied voltage does. The current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage, so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle the current is flowing in the normal direction through the circuit, since the capacitor is being charged. Hence the current is positive, as indicated by the dashed line in Fig. 2-26.

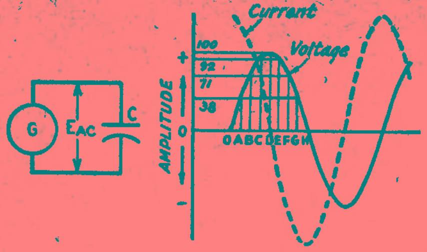


Fig. 2-26 — Voltage and current phase relationships when an alternating voltage is applied to a capacitor.

In the second quarter cycle – that is, in the time from *D* to *H*, the voltage applied to the capacitor decreases. During this time the capacitor *loses* its charge. Applying the same reasoning, it is plain that the current is small in interval *DE* and continues to increase during each succeeding interval. However, the current is flowing *against* the applied voltage because the capacitor is discharging into the circuit. The current flows in the *negative* direction during this quarter cycle.

The third and fourth quarter cycles repeat the events of the first and second, respectively, with this difference – the polarity of the applied voltage has reversed, and the current changes to correspond. In other words, an alternating current flows in the circuit because of the alternate charging and discharging of the capacitance. As shown by Fig. 2-26, the current starts its cycle 90 degrees before the voltage, so the current in a capacitor leads the applied voltage by 90 degrees.

Capacitive Reactance

The quantity of electric charge that can be placed on a capacitor is proportional to the applied emf and the capacitance. This amount of charge moves back and forth in the circuit once each cycle, and so the *rate* of movement of charge – that is, the current – is proportional to voltage, capacitance and frequency. If the effects of capacitance and frequency are lumped together, they form a quantity that plays a part similar to that of resistance in Ohm's Law. This quantity is called reactance, and the unit for it is the ohm, just as in the case of resistance. The formula for it is

$$X_C = \frac{1}{2\pi fC}$$

where X_C = Capacitive reactance in ohms
f = Frequency in cycles per second
C = Capacitance in farads
π = 3.14

Although the unit of reactance is the ohm, there is no power dissipation in reactance. The energy stored in the capacitor in one quarter of the cycle is simply returned to the circuit in the next.

The fundamental units (cycles per second, farads) are too large for practical use in radio circuits. However, if the capacitance is in microfarads and the frequency is in megacycles, the reactance will come out in ohms in the formula.

Example: The reactance of a capacitor of 470 pF (0.00047 μF) at a frequency of 7150 kHz (7.15 MHz) is

$$X = \frac{1}{2\pi fC} = \frac{1}{6.28 \times 7.15 \times .00047} = 47.4 \text{ ohms}$$

Inductive Reactance

When an alternating voltage is applied to a *pure* inductance (one with no resistance – all *practical* inductors have resistance) the current is again 90 degrees out of phase with the applied voltage. However, in this case the current *lags* 90 degrees behind the voltage – the opposite of the capacitor current-voltage relationship.

The primary cause for this is the *back emf* generated in the inductance, and since the amplitude of the back emf is proportional to the rate at which the current changes, and this in turn is proportional to the frequency, the amplitude of the current is inversely proportional to the applied frequency. Also, since the back emf is proportional to inductance for a given rate of current change, the current flow is inversely proportional to inductance for a given applied voltage and frequency. (Another way of saying this is that just enough current flows to generate an induced emf that equals and opposes the applied voltage.)

The combined effect of inductance and frequency is called inductive reactance, also expressed in ohms, and the formula for it is

$$X_L = 2\pi fL$$

where X_L = Inductive reactance in ohms
f = Frequency in cycles per second
L = Inductance in henrys
π = 3.14

Example: The reactance of a 15-microhenry coil at a frequency of 14 MHz is

$$X_L = 2\pi fL = 6.28 \times 14 \times 15 = 1319 \text{ ohms}$$

In radio-frequency circuits the inductance values usually are small and the frequencies are large. If the inductance is expressed in millihenrys and the frequency in kilocycles, the conversion factors for the two units cancel, and the formula for reactance may be used without first converting to fundamental units. Similarly, no conversion is necessary if the inductance is in microhenrys and the frequency is in megacycles.

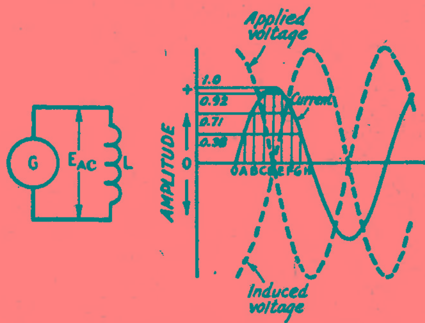


Fig. 2-27 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

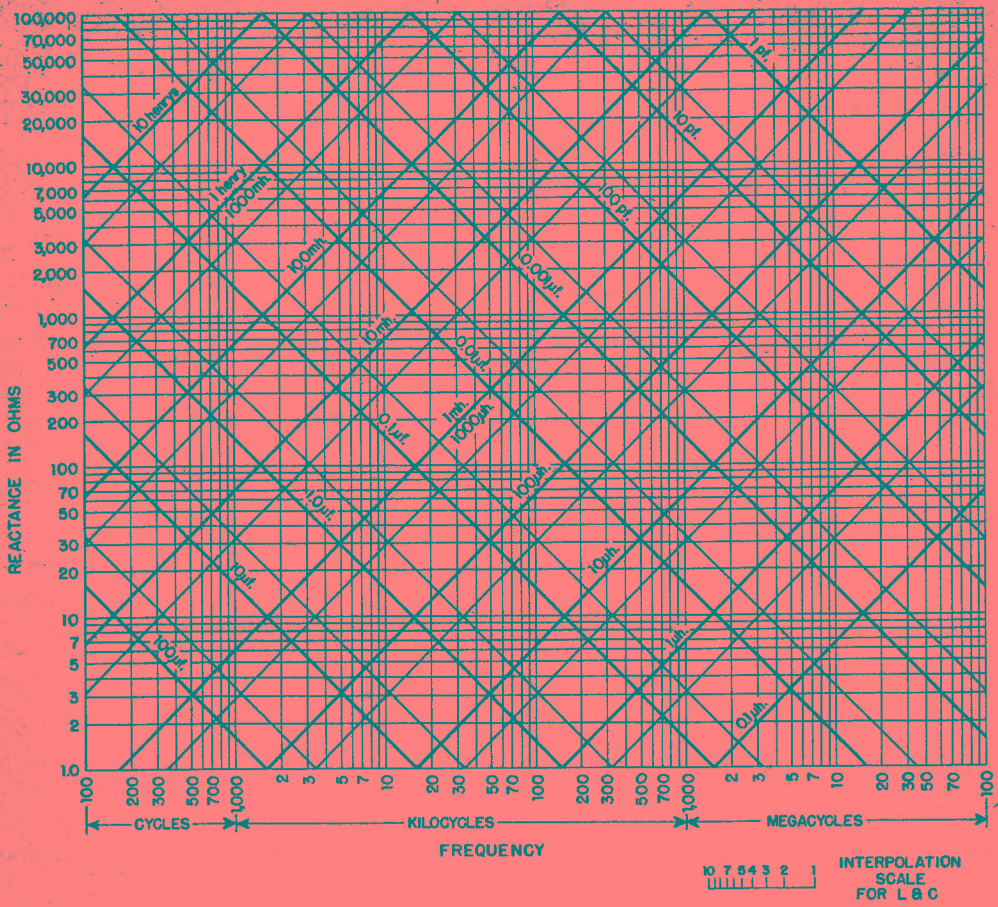


Fig. 2-28 — Inductive and capacitive reactance vs. frequency. Heavy lines represent multiples of 10, intermediate light lines multiples of 5; e.g., the light line between 10 µH and 100 µH represents 50 µH, the light line between 0.1 µF and 1 µF represents 0.5 µF, etc. Intermediate values can be estimated with the help of the interpolation scale.

Reactances outside the range of the chart may be found by applying appropriate factors to values within the chart range. For example, the reactance of 10 henrys at 60 cycles can be found by taking the reactance to 10 henrys at 600 cycles and dividing by 10 for the 10-times decrease in frequency.

Example: The reactance of a coil having an inductance of 8 henrys, at a frequency of 120 cycles, is

$$X_L = 2\pi fL = 6.28 \times 120 \times 8 = 6029 \text{ ohms}$$

The resistance of the wire of which the coil is wound has no effect on the reactance, but simply acts as though it were a separate resistor connected in series with the coil.

Ohm's Law for Reactance

Ohm's Law for an ac circuit containing *only* reactance is

$$I = \frac{E}{X}$$

$$E = IX$$

$$X = \frac{E}{I}$$

where E = Emf in volts
I = Current in amperes
X = Reactance in ohms

The reactance in the circuit may, of course, be either inductive or capacitive.

Example: If a current of 2 amperes is flowing through the capacitor of the earlier example (reactance = 47.4 ohms) at 7150 kHz, the voltage drop across the capacitor is

$$E = IX = 2 \times 47.4 = 94.8 \text{ volts}$$

If 400 volts at 120 hertz is applied to the 8-henry inductor of the earlier example, the current through the coil will be

$$I = \frac{E}{X} = \frac{420}{6029} = 0.0663 \text{ amp. (66.3 mA)}$$

Reactance Chart

The accompanying chart, Fig. 2-28, shows the reactance of capacitances from 1 pF to 100 µF, and the reactance of inductances from 0.1 µH to 10 henrys, for frequencies between 100 hertz and 100 megahertz per second. The approximate value of reactance can be read from the chart or, where more exact values are needed, the chart will serve as a check on the order of magnitude of reactances calculated from the formulas given above, and thus avoid "decimal-point errors."

Reactances in Series and Parallel

When reactances of the same kind are connected in series or parallel the resultant reactance is that of the resultant inductance or capacitance. This leads to the same rules that are used when determining the resultant resistance when resistors are combined. That is, for series reactances of the same kind the resultant reactance is

$$X = X1 + X2 + X3 + X4$$

and for reactances of the same kind in parallel the resultant is

$$X = \frac{1}{\frac{1}{X1} + \frac{1}{X2} + \frac{1}{X3} + \frac{1}{X4}}$$

or for two in parallel,

$$X = \frac{X1X2}{X1 + X2}$$

The situation is different when reactances of opposite kinds are combined. Since the current in a capacitance leads the applied voltage by 90 degrees and the current in an inductance lags the applied voltage by 90 degrees, the voltages at the terminals of opposite types of reactance are 180 degrees out of phase in a series circuit (in which the current has to be the same through all elements), and the currents in reactances of opposite types are 180 degrees out of phase in a parallel circuit (in which the same voltage is applied to all elements). The 180-degree phase relationship means that the currents or voltages are of opposite polarity, so in the series circuit of Fig. 2-29A the voltage EL across the inductive reactance XL is of opposite polarity to the voltage EC across the capacitive reactance XC. Thus if we call XL "positive" and XC "negative" (a common convention) the applied voltage EAC is $EL - EC$. In the parallel circuit at B the total current, I, is equal to $IL - IC$, since the currents are 180 degrees out of phase.

In the series case, therefore, the resultant reactance of XL and XC is

$$X = X_L - X_C$$

and in the parallel case

$$X = \frac{-X_L X_C}{X_L - X_C}$$

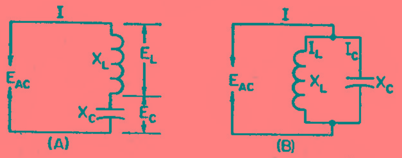


Fig. 2-29 — Series and parallel circuits containing opposite kinds of reactance.

Note that in the series circuit the total reactance is negative if XC is larger than XL; this indicates that the total reactance is capacitive in such a case. The resultant reactance in a series circuit is always smaller than the larger of the two individual reactances.

In the parallel circuit, the resultant reactance is negative (i.e., capacitive) if XL is larger than XC, and positive (inductive) if XL is smaller than XC, but in every case is always larger than the smaller of the two individual reactances.

In the special case where $XL = XC$ the total reactance is zero in the series circuit and infinitely large in the parallel circuit.

Reactive Power

In Fig. 2-29A the voltage drop across the inductor is larger than the voltage applied to the circuit. This might seem to be an impossible condition, but it is not; the explanation is that while energy is being stored in the inductor's magnetic field, energy is being returned to the circuit from the capacitor's electric field, and vice versa. This stored energy is responsible for the fact that the voltages across reactances in series can be larger than the voltage applied to them.

In a resistance the flow of current causes heating and a power loss equal to I^2R. The power in a reactance is equal to I^2X, but is not a "loss"; it is simply power that is transferred back and forth between the field and the circuit but not used up in heating anything. To distinguish this "nondissipated" power from the power which is actually consumed, the unit of reactive power is called the volt-ampere-reactive, or var, instead of the watt. Reactive power is sometimes called "wattless" power.

IMPEDANCE

When a circuit contains both resistance and reactance the combined effect of the two is called impedance, symbolized by the letter Z. (Impedance is thus a more general term than either resistance or reactance, and is frequently used even for circuits that have only resistance or reactance, although usually with a qualification – such as "resistive impedance" to indicate that the circuit has only resistance, for example.)

The reactance and resistance comprising an impedance may be connected either in series or in parallel, as shown in Fig. 2-30. In these circuits the reactance is shown as a box to indicate that it may be either inductive or capacitive. In the series circuit the current is the same in both elements, with (generally) different voltages appearing across the resistance and reactance. In the parallel circuit the same voltage is applied to both elements, but different currents flow in the two branches.

Since in a resistance the current is in phase with the applied voltage while in a reactance it is 90 degrees out of phase with the voltage, the phase relationship between current and voltage in the circuit as a whole may be anything between zero and 90 degrees, depending on the relative amounts of resistance and reactance.

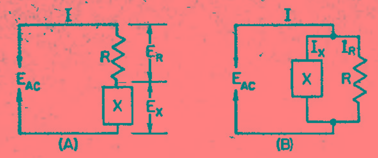


Fig. 2-30 — Series and parallel circuits containing resistance and reactance.

Series Circuits

When resistance and reactance are in series, the impedance of the circuit is

$$Z = \sqrt{R^2 + X^2}$$

where Z = Impedance in ohms
R = Resistance in ohms
X = Reactance in ohms

The reactance may be either capacitive or inductive. If there are two or more reactances in the circuit they may be combined into a resultant by the rules previously given, before substitution into the formula above; similarly for resistances.

The "square root of the sum of the squares" rule for finding impedance in a series circuit arises from the fact that the voltage drops across the resistance and reactance are 90 degrees out of phase, and so combine by the same rule that applies in finding the hypothenuse of a right-angled triangle when the base and altitude are known.

Parallel Circuits

With resistance and reactance in parallel, as in Fig. 2-30B, the impedance is

$$Z = \frac{RX}{\sqrt{R^2 + X^2}}$$

where the symbols have the same meaning as for series circuits.

Just as in the case of series circuits, a number of reactances in parallel should be combined to find the resultant reactance before substitution into the formal above; similarly for a number of resistances in parallel.

Equivalent Series and Parallel Circuits

The two circuits shown in Fig. 2-30 are equivalent if the same current flows when a given voltage of the same frequency is applied, and if the phase angle between voltage and current is the same in both cases. It is in fact possible to "transform" any given series circuit into an equivalent parallel circuit, and vice versa.

Transformations of this type often lead to simplification in the solution of complicated circuits. However, from the standpoint of practical work the usefulness of such transformations lies in the fact that the impedance of a circuit may be modified by the addition of *either* series or parallel elements, depending on which happens to be most convenient in the particular case. Typical applications are considered later in connection with tuned circuits and transmission lines.

Ohm's Law for Impedance

Ohm's Law can be applied to circuits containing impedance just as readily as to circuits having resistance or reactance only. The formulas are

$$I = \frac{E}{Z}$$

$$E = IZ$$

$$Z = \frac{E}{I}$$

where E = Emf in volts
I = Current in amperes
Z = Impedance in ohms

Fig. 2-31 shows a simple circuit consisting of a resistance of 75 ohms and a reactance of 100 ohms in series. From the formula previously given, the impedance is

$$Z = \sqrt{R^2 + X_L^2} = \sqrt{(75)^2 + (100)^2} = 125$$

If the applied voltage is 250 volts, then

$$I = \frac{E}{Z} = \frac{250}{125} = 2 \text{ amperes}$$

This current flows through both the resistance and reactance, so the voltage drops are

$$E_R = IR = 2 \times 75 = 150 \text{ volts}$$
$$E_{XL} = IX_L = 2 \times 100 = 200 \text{ volts}$$

The simple arithmetical sum of these two drops, 350 volts, is greater than the applied voltage because the two voltages are 90 degrees out of phase. Their actual resultant, when phase is taken into account, is

$$\sqrt{(150)^2 + (200)^2} = 250 \text{ volts}$$

Power Factor

In the circuit of Fig. 2-31 an applied emf of 250 volts results in a current of 2 amperes, giving an apparent power of 250 X 2 = 500 watts. However, only the resistance actually consumes power. The power in the resistance is

$$P = I^2R = (2)^2 \times 75 = 300 \text{ watts}$$

The ratio of the power consumed to the apparent power is called the power factor of the circuit, and in this example the power factor would be 300/500 = 0.6. Power factor is frequently expressed as a percentage; in this case, it would be 60 percent.

"Real" or dissipated power is measured in watts; apparent power, to distinguish it from real power, is measured in volt-amperes. It is simply the product of volts and amperes and has no direct relationship to the power actually used up or dissipated unless the power factor of the circuit is known. The power factor of a purely resistive circuit is 100 percent or 1, while the power factor of a pure reactance is zero. In this illustration, the reactive power is $VAR = I2X = (2)2 \times 100 = 400$ volt-amperes.

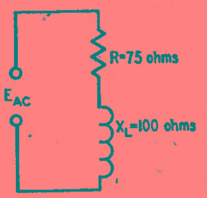


Fig. 2-31 — Circuit used as an example for impedance calculations.

Reactance and Complex Waves

It was pointed out earlier in this chapter that a complex wave (a "nonsinusoidal" wave) can be resolved into a fundamental frequency and a series of harmonic frequencies. When such a complex voltage wave is applied to a circuit containing reactance, the current through the circuit will not have the same wave shape as the applied voltage. This is because the reactance of an inductor and capacitor depend upon the applied frequency. For the second-harmonic component of a complex wave, the reactance of the inductor is twice and the reactance of the capacitor one-half their respective values at the fundamental frequency; for the third harmonic the inductor reactance is three times and the capacitor reactance one-third, and so on. Thus the circuit impedance is different for each harmonic component.

Just what happens to the current wave shape depends upon the values of resistance and reactance involved and how the circuit is arranged. In a simple circuit with resistance and inductive reactance in series, the amplitudes of the harmonic currents will be reduced because the inductive reactance increases in proportion to frequency. When capacitance and resistance are in series, the harmonic current is likely to be accentuated because the capacitive reactance becomes lower as the frequency is raised. When both inductive and capacitive reactance are present the shape of the current wave can be altered in a variety of ways, depending upon the circuit and the "constants," or the relative values of *L*, *C*, and *R*, selected.

This property of nonuniform behavior with respect to fundamental and harmonics is an extremely useful one. It is the basis of "filtering," or the suppression of undesired frequencies in favor of a single desired frequency or group of such frequencies.

TRANSFORMERS FOR AUDIO FREQUENCIES

Two coils having mutual inductance constitute a **transformer**. The coil connected to the source of energy is called the **primary** coil, and the other is called the **secondary** coil.

The usefulness of the transformer lies in the fact that electrical energy can be transferred from one circuit to another without direct connection, and in the process can be readily changed from one voltage level to another. Thus, if a device to be operated requires, for example, 115 volts ac and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. A transformer can be used only with ac, since no voltage will be induced in the secondary if the magnetic field is not changing. If dc is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

THE IRON-CORE TRANSFORMER

As shown in Fig. 2-32, the primary and secondary coils of a transformer may be wound on a core of magnetic material. This increases the inductance of the coils so that a relatively small number of turns may be used to induce a given value of voltage with a small current. A closed core (one having a continuous magnetic path) such as that shown in Fig. 2-32 also tends to insure that practically all of the field set up by the current in the primary coil will cut the turns of the secondary coil. However, the core introduces a power loss because of hysteresis and eddy currents so this type of construction is normally practicable only at power and audio frequencies. The discussion in this section is confined to transformers operating at such frequencies.

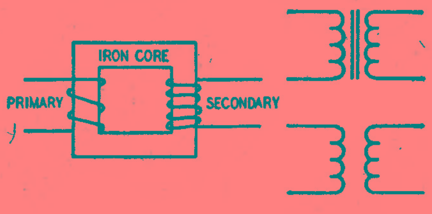


Fig. 2-32 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

Voltage and Turns Ratio

For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns in the coil. If the two coils of a transformer are in the same field (which is the case when both are wound on the same closed core) it follows that the induced voltages will be proportional to the number of turns in each coil. In the primary the induced voltage is practically equal to, and opposes, the applied voltage, as described earlier. Hence,

$$E_s = \frac{n_s}{n_p} E_p$$

where E_s = Secondary voltage
E_p = Primary applied voltage
n_s = Number of turns on secondary
n_p = Number of turns on primary

The ratio, ns/np is called the secondary-to-primary turns ratio of the transformer.

Example: A transformer has a primary of 400 turns and a secondary of 2800 turns, and an emf of 115 volts is applied to the primary.

$$E_s = \frac{n_s}{n_p} E_p = \frac{2800}{400} \times 115 = 7 \times 115$$

$$= 805 \text{ volts}$$

Also, if an emf of 805 volts is applied to the 2800-turn winding (which then becomes the primary) the output voltage from the 400-turn winding will be 115 volts.

Either winding of a transformer can be used as the primary, providing the winding has enough turns (enough inductance) to induce a voltage equal to the applied voltage without requiring an excessive current flow.

Effect of Secondary Current

The current that flows in the primary when no current is taken from the secondary is called the **magnetizing current** of the transformer. In any properly-designed transformer the primary inductance will be so large that the magnetizing current will be quite small. The power consumed by the transformer when the secondary is "open" – that is, not delivering power – is only the amount necessary to supply the losses in the iron core and in the resistance of the wire with which the primary is wound.

When power is taken from the secondary winding, the secondary current sets up a magnetic field that opposes the field set up by the primary current. But if the induced voltage in the primary is to equal the applied voltage, the original field must be maintained. Consequently, the primary must draw enough additional current to set up a field exactly equal and opposite to the field set up by the secondary current.

In practical calculations on transformers it may be assumed that the entire primary current is caused by the secondary "load." This is justifiable because the magnetizing current should be very small in comparison with the primary "load" current at rated power output.

If the magnetic fields set up by the primary and secondary currents are to be equal, the primary current multiplied by the primary turns must equal the secondary current multiplied by the secondary turns. From this it follows that

$$I_p = \frac{n_s}{n_p} I_s$$

where I_p = Primary current
I_s = Secondary current
n_p = Number of turns on primary
n_s = Number of turns on secondary

Example: Suppose that the secondary of the transformer in the previous example is delivering a current of 0.2 ampere to a load. Then the primary current will be

$$I_p = \frac{n_s}{n_p} I_s = \frac{2800}{400} \times 0.2 = 7 \times 0.2 = 1.4 \text{ amp.}$$

Although the secondary voltage is higher than the primary voltage, the secondary *current* is *lower* than the primary current, and by the same ratio.

Power Relationships; Efficiency

A transformer cannot create power; it can only transfer it and change the emf. Hence, the power taken from the secondary cannot exceed that taken by the primary from the source of applied emf. There is always some power loss in the resistance of the coils and in the iron core, so in all practical cases the power taken from the source will exceed that taken from the secondary. Thus,

$$P_o = nP_i$$

where P_o = Power output from secondary
P_i = Power input to primary
n = Efficiency factor

The efficiency, n, always is less than 1. It is usually expressed as a percentage; if n is 0.65, for instances, the efficiency is 65 percent.

Example: A transformer has an efficiency of 85 percent at its full-load output of 150 watts. The power input to the primary at full secondary load will be

$$P_i = \frac{P_o}{n} = \frac{150}{0.85} = 176.5 \text{ watts}$$

A transformer is usually designed to have its highest efficiency at the power output for which it is rated. The efficiency decreases with either lower or higher outputs. On the other hand, the *losses* in the transformer are relatively small at low output but increase as more power is taken. The amount of power that the transformer can handle is determined by its own losses, because these heat the wire and core. There is a limit to the temperature rise that can be tolerated, because too-high temperature either will melt the wire or cause the insulation to break down. A transformer can be operated a reduced output, even though the efficiency is low, because the actual loss will be low under such conditions.

The full-load efficiency of small power transformers such as are used in radio receivers and transmitters usually lies between about 60 and 90 percent, depending upon the size and design.

Leakage Reactance

In a practical transformer not all of the magnetic flux is common to both windings, although in well-designed transformers the amount of flux that "cuts" one coil and not the other is only a small percentage of the total flux. This **leakage flux** causes an emf of self-induction; consequently, there are small amounts of **leakage inductance** associated with both windings of the transformer. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of leakage inductance and the frequency. This reactance is called **leakage reactance**.

Current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing current, hence it increases as more power is taken from the secondary. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistances of the transformer windings also cause voltage drops when current is flowing; although these voltage drops are not in phase with those caused by leakage reactance, together they result in a lower secondary voltage under load than is indicated by the turns ratio of the transformer.

At power frequencies (60 cycles) the voltage at the secondary, with a reasonably well-designed transformer, should not drop more than about 10 percent from open-circuit conditions to full load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies because the leakage reactance increases directly with the frequency.

Impedance Ratio

In an ideal transformer – one without losses or leakage reactance – the following relationship is true:

$$Z_p = Z_s \left[\frac{N_p}{N_s}\right]^2$$

where Z_p = Impedance looking into primary terminals from source of power
Z_s Impedance of load connected to secondary
N_p/N_s = Turns ratio, primary to secondary

That is, a load of any given impedance connected to the secondary of the transformer will be transformed to a different value "looking into" the primary from the source of power. The impedance transformation is proportional to the square of the primary-to-secondary turns ratio.

Example: A transformer has a primary-to-secondary turns ratio of 0.6 (primary has 6/10 as many turns as the secondary) and a load of 3000 ohms is connected to the secondary. The impedance looking into the primary then will be

$$Z_p = Z_s \left[\frac{N_p}{N_s}\right] = 3000 \times (0.6)^2 = 3000 \times 0.36$$
$$= 1080 \text{ ohms}$$

By choosing the proper turns ratio, the impedance of a fixed load can be transformed to any desired value, within practical limits. If transformer losses can be neglected, the transformed or "reflected" impedance has the same phase angle as the actual load impedance; thus if the load is a pure resistance the load presented by the primary to the source of power also will be a pure resistance.

The above relationship may be used in practical work even though it is based on an "ideal" transformer. Aside from the normal design requirements of reasonably low internal losses and low leakage reactance, the only requirement is that the primary have enough inductance to operate with low magnetizing current at the voltage applied to the primary.

The primary impedance of a transformer – *as it appears to the source of power* – is determined wholly by the load connected to the secondary and by the turns ratio. If the characteristics of the transformer have an appreciable effect on the impedance presented to the power source, the transformer is either poorly designed or is not suited to the voltage and frequency at which it is being used. Most transformers will operate quite well at voltages from slightly above to well below the design figure.

Impedance Matching

Many devices require a specific value of load resistance (or impedance) for optimum operation.

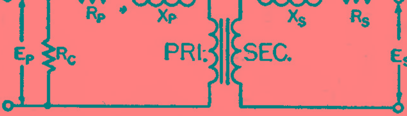


Fig. 2-33 – The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondery windings. The resistance *Rc* is an equivalent resistance representing the core losses, which are essentially constant for any given applied voltage and frequency. Since these ere comparatively small, their effect may be neglected in many approximate calculations.

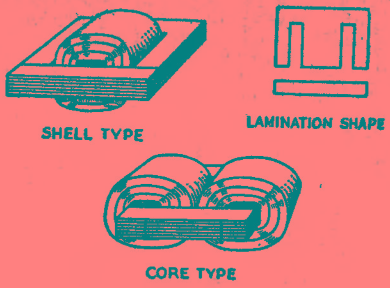


Fig. 2-34 – Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path.

The impedance of the actual load that is to dissipate the power may differ widely from this value, so a transformer is used to change the actual load into an impedance of the desired value. This is called impedance matching. From the preceding,

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}}$$

where N_p/N_s = Required turns ratio, primary to secondary
Z_p = Primary impedance required
Z_s = Impedance of load connected to secondary

Example: A vacuum-tube af amplifier requires a load of 5000 ohms for optimum performance, and is to be connected to a loud-speaker having an impedance of 10 ohms. The turns ratio, primary to secondary, required in the coupling transformer is

$$\frac{N_p}{N_s} = \sqrt{\frac{Z_p}{Z_s}} = \sqrt{\frac{5000}{10}} = \sqrt{500} = 22.4$$

The primary therefore must have 22.4 times as many turns as the secondary.

Impedance matching means, in general, adjusting the load impedance – by means of a transformer or otherwise – to a desired value. However, there is also another meaning. It is possible to show that any source of power will deliver its maximum possible output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. The efficiency is only 50 percent in such a case; just as much power is used up in the source as is delivered to the load. Because of the poor efficiency, this type of impedance matching is limited to cases where only a small amount of power is available and heating from power loss in the source is not important.

Transformer Construction

Transformers usually are designed so that the magnetic path around the core is as short as possible. A short magnetic path means that the transformer will operate with fewer turns, for a given applied voltage, than if the path were long. A short path also helps to reduce flux leakage and therefore minimizes leakage reactance.

Two core shapes are in common use, as shown in Fig. 2-34. In the shell type both windings are placed on the inner leg, while in the core type the

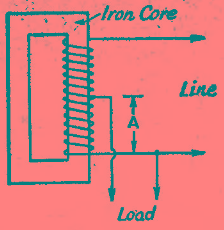


Fig. 2-35 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding *(A)* flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

primary and secondary windings may be placed on separate legs, if desired. This is sometimes done when it is necessary to minimize capacitive effects between the primary and secondary, or when one of the windings must operate at very high voltage.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of laminations, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of eddy currents. The laminations are interleaved at the ends to make the magnetic path as continuous as possible and thus reduce flux leakage.

The number of turns required in the primary for a given applied emf is determined by the size, shape and type of core material used, and the frequency. The number of turns required is inversely proportional to the cross-sectional area of the core. As a rough indication, windings of small power transformers frequently have about six to eight turns per volt on a core of 1-square-inch cross section and have a magnetic path 10 or 12 inches in length. A longer path or smaller cross section requires more turns per volt, and vice versa.

In most transformers the coils are wound in layers, with a thin sheet of treated-paper insulation between each layer. Thicker insulation is used between coils and between coils and core.

Autotransformers

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 2-35; the principles just discussed apply equally well. A one-winding transformer is called an autotransformer. The current in the common section (A) of the winding is the difference between the line (primary) and the load (secondary) currents, since these currents are out of phase. Hence if the line and load currents are nearly equal the common section of the winding may be wound with comparatively small wire. This will be the case only when the primary (line) and secondary (load) voltages are not very different. The autotransformer is used chiefly for boosting or reducing the power-line voltage by relatively small amounts. Continuously-variable autotransformers are commercially available under a variety of trade names; "Variac" and "Powerstat" are typical examples.

THE DECIBEL

In most radio communication the received signal is converted into sound. This being the case, it is useful to appraise signal strengths in terms of relative loudness as registered by the ear. A peculiarity of the ear is that an increase or decrease in loudness is responsive to the *ratio* of the amounts of power involved, and is practically independent of absolute value of the power. For example, if a person estimates that the signal is "twice as loud" when the transmitter power is increased from 10 watts to 40 watts, he will also estimate that a 400-watt signal is twice as loud as a 100-watt signal. In other words, the human ear has a *logarithmic* response.

This fact is the basis for the use of the relative-power unit called the decibel (abbreviated dB). A change of one decibel in the power level is just detectable as a change in loudness under ideal conditions. The number of decibels corresponding to a given power ratio is given by the following formula:

$$dB = 10 \log \frac{P_2}{P_1}$$

Common logarithms (base 10) are used.

Voltage and Current Ratios

Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but only when the impedance is the same for both values of voltage, or current. The gain of an amplifier cannot be expressed correctly in dB if it is based on the ratio of the output voltage to the input voltage unless both voltages are measured across the same value of impedance. When the impedance at both points of measurement is the same, the following formula may be used for voltage or current ratios:

$$dB = 20 \log \frac{V_2}{V_1} \quad \text{or } 20 \log \frac{I_2}{I_1}$$

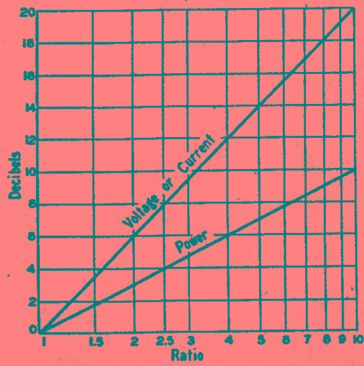


Fig. 2-36 — Decibel chart for power, voltage and current ratios for power ratios of 1:1 and 10:1. In determining decibels for current or voltage ratios the currents (or voltages) being compared must be referred to the same value of impedance.

Decibel Chart

The two formulas are shown graphically in Fig. 2-36 for ratios from 1 to 10. Gains (increases) expressed in decibels may be added arithmetically; losses (decreases) may be subtracted. A power decrease is indicated by prefixing the decibel figure with a minus sign. Thus +6 dB means that the power has been multiplied by 4, while –6 dB means that the power has been divided by 4.

The chart may be used for other ratios by adding (or subtracting, if a loss) 10 dB each time the ratio scale is multiplied by 10, for power ratios; or by adding (or subtracting) 20 dB each time the scale is multiplied by 10 for voltage or current ratios. For example, a power ratio of 2.5 is 4 dB (from the chart). A power ratio of 10 times 2.5, or 25, is 14 dB (10 + 4), and a power ratio of 100 times 2.5, or 250, is 24 dB (20 + 4). A voltage or current ratio of 4 is 12 dB, a voltage or current ratio of 40 is 32 dB (20 + 12), and one of 400 is 52 dB (40 + 12).

RADIO-FREQUENCY CIRCUITS

RESONANCE IN SERIES CIRCUITS

Fig. 2-37 shows a resistor, capacitor and inductor connected in series with a source of alternating current, the frequency of which can be varied over a wide range. At some *low* frequency the capacitive reactance will be much larger than the resistance of *R*, and the inductive reactance will be small compared with either the reactance of *C* or the resistance of *R*. (*R* is assumed to be the same at all frequencies.) On the other hand, at some very *high* frequency the reactance of *C* will be very small and the reactance of *L* will be very large. In either case the current will be small, because the net reactance is large.

At some intermediate frequency, the reactances of *C* and *L* will be equal and the voltage drops across the coil and capacitor will be equal and 180 degrees out of phase. Therefore they cancel each other completely and the current flow is determined wholly by the resistance, *R*. At that frequency the current has its largest possible value, assuming the source voltage to be constant regardless of frequency. A series circuit in which the inductive and capacitive reactances are equal is said to be resonant.

The principle of resonance finds its most extensive application in radio-frequency circuits. The reactive effects associated with even small inductances and capacitances would place drastic limitations on rf circuit operation if it were not possible to "cancel them out" by supplying the right amount of reactance of the opposite kind – in other words, "tuning the circuit to resonance."

Resonant Frequency

The frequency at which a series circuit is resonant is that for which $XL = XC$. Substituting the formulas for inductive and capacitive reactance gives

$$f = \frac{1}{2\pi\sqrt{LC}}$$

where f = Frequency in cycles per second
L = Inductance in henrys
C = Capacitance in farads
π = 3.14

These units are inconveniently large for radio-frequency circuits. A formula using more appropriate units is

$$f = \frac{10^6}{2\pi\sqrt{LC}}$$

where f = Frequency in kilohertz (kHz)
L = Inductance in microhenrys (μH)
C = Capacitance in picofarads (pF)
π = 3.14

Example: The resonant frequency of a series circuit containing a 5-μH inductor and a 35-pF capacitor is

$$f = \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}}$$

$$= \frac{10^6}{6.28 \times 13.2} = \frac{10^6}{83} = 12{,}050 \text{ kHz}$$

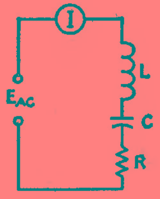


Fig. 2-37 – A series circuit containing *L*, *C* and *R* is "resonant" at the applied frequency when the reactance of *C* is equal to the reactance of *L*.

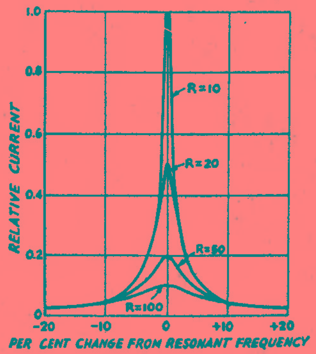


Fig. 2-38 – Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms. Note that at frequencies more than plus or minus ten percent away from the resonant frequency the current is substantially uneffected by the resistance in the circuit.

The formula for resonant frequency is not affected by resistance in the circuit.

Resonance Curves

If a plot is drawn on the current flowing in the circuit of Fig. 2-37 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-38. The shape of the resonance curve at frequencies near resonance is determined by the ratio of reactance to resistance.

If the reactance of either the coil or capacitor is of the same order of magnitude as the resistance, the current decreases rather slowly as the frequency is moved in either direction away from resonance. Such a curve is said to be broad. On the other hand, if the reactance is considerably larger than the resistance the current decreases rapidly as the frequency moves away from resonance and the circuit is said to be sharp. A sharp circuit will respond a great deal more readily to the resonant frequency than to frequencies quite close to resonance; a broad circuit will respond almost equally well to a group or band of frequencies centering around the resonant frequency.

Both types of resonance curves are useful. A sharp circuit gives good selectivity – the ability to respond strongly (in terms of current amplitude) at one desired frequency and discriminate against others. A broad circuit is used when the apparatus must give about the same response over a band of frequencies rather than to a single frequency alone.

Q

Most diagrams of resonant circuits show only inductance and capacitance; no resistance is indicated. Nevertheless, resistance is always present. At frequencies up to perhaps 30 MHz this resistance is mostly in the wire of the coil. Above this frequency energy loss in the capacitor (principally in the solid dielectric which must be used to form an insulating support for the capacitor plates) also becomes a factor. This energy loss is equivalent to resistance. When maximum sharpness or selectivity is needed the object of design is to reduce the inherent resistance to the lowest possible value.

The value of the reactance of either the inductor or capacitor at the resonant frequency of a series-resonant circuit, divided by the *series* resistance in the circuit, is called the Q (quality factor) of the circuit, or

$$Q = \frac{X}{r}$$

where Q = Quality factor
X = Reactance of either coil or capacitor in ohms
r = Series resistance in ohms

Example: The inductor and capacitor in a series circuit each have a reactance of 350 ohms at the resonant frequency. The resistance is 5 ohms. Then the Q is

$$Q = \frac{X}{r} = \frac{350}{5} = 70$$

The effect of Q on the sharpness of resonance of a circuit is shown by the curves of Fig. 2-39. In these curves the frequency change is shown in percentage above and below the resonant frequency. Qs of 10, 20, 50 and 100 are shown; these values cover much of the range commonly used in radio work. The unloaded Q of a circuit is determined by the inherent resistances associated with the components.

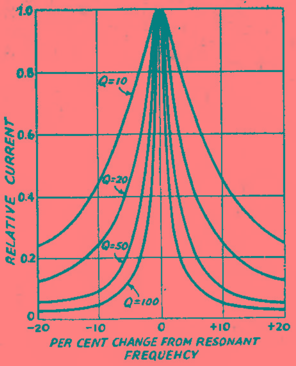


Fig. 2-39 — Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

Voltage Rise at Resonance

When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage that appears across either the inductor or capacitor is considerably higher than the applied voltage. The current in the circuit is limited only by the resistance and may have a relatively high value; however, the same current flows through the high reactances of the inductor and capacitor and causes large voltage drops. The ratio of the reactive voltage to the applied voltage is equal to the ratio of reactance to resistance. This ratio is also the Q of the circuit. Therefore, the voltage across either the inductor or capacitor is equal to QE where E is the voltage inserted in series. This fact accounts for the high voltages developed across the components of series-tuned antenna couplers (see chapter on "Transmission Lines").

RESONANCE IN PARALLEL CIRCUITS

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-40 there is a resonance effect similar to that in a series circuit. However, in this case the "line" current (measured at the point indicated) is *smallest* at the frequency for which the inductive and capacitive reactances are equal. At that frequency the current through L is exactly canceled by the out-of-phase current through C, so that only the current taken by R flows in the line. At frequencies *below* resonance the current through L is larger than that through C, because the reactance of L is smaller and that of C higher at low frequencies; there is only partial cancellation of the two reactive currents and the line current therefore is larger than the current taken by R alone. At frequencies *above* resonance the situation is reversed and more current flows through C than

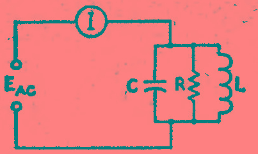


Fig. 2-40 — Circuit illustrating parallel resonance.

through L, so the line current again increases. The current at resonance, being determined wholly by R, will be small if R is large and large if R is small.

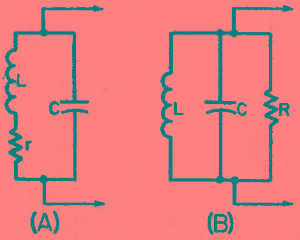
The resistance R shown in Fig. 2-40 is not necessarily an actual resistor. In many cases it will be the series resistance of the coil "transformed" to an equivalent parallel resistance (see later). It may be antenna or other load resistance coupled into the tuned circuit. In all cases it represents the total effective resistance in the circuit.

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-41 will behave identically, when an external voltage is applied, if (1) L and C are the same in both cases; and (2) R multiplied by r, equals the square of the reactance (at resonance) of either L or C. When these conditions are met the two circuits will have the same Q. (These statements are approximate, but are quite accurate if the Q is 10 or more.) The circuit at A is a *series* circuit if it is viewed from the "inside" — that is, going around the loop formed by L, C and r — so its Q can be found from the ratio of X to r.

Thus a circuit like that of Fig. 2-41A has an equivalent parallel impedance (at resonance) of $R = \frac{X^2}{r}$; X is the reactance of either the inductor or the capacitor. Although R is not an actual resistor, to the source of voltage the parallel-resonant circuit "looks like" a pure resistance of that value. It is "pure" resistance because the inductive and capacitive currents are 180 degrees out of phase and are equal; thus there is no reactive current in the line. In a practical circuit with a high-Q capacitor, at the resonant frequency the parallel impedance is

$$Z_r = QX$$

where Z_r = Resistive impedance at resonance
Q = Quality factor of inductor
X = Reactance (in ohms) of either the inductor or capacitor



(A) (B)

Fig. 2-41 — Series and parallel equivalents when the two circuits are resonant. The series resistance, r, in A is replaced in B by the equivalent parallel resistance ($R = X^2c/r = X^2L/r$) and vice versa.

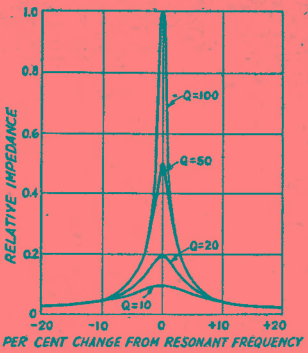


Fig. 2-42 — Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-39 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

Example: The parallel impedance of a circuit with a coil Q of 50 and having inductive and capacitive reactance of 300 ohms will be

$$Z_r = QX = 50 \times 300 = 15{,}000 \text{ ohms}$$

At frequencies off resonance the impedance is no longer purely resistive because the inductive and capacitive currents are not equal. The off-resonant impedance therefore is complex, and is lower than the resonant impedance for the reasons previously outlined.

The higher the Q of the circuit, the higher the parallel impedance. Curves showing the variation of impedance (with frequency) of a parallel circuit have just the same shape as the curves showing the variation of current with frequency in a series circuit. Fig. 2-42 is a set of such curves. A set of curves showing the relative response as a function of the departure from the resonant frequency would be similar to Fig. 2-39. The −3 dB bandwidth (bandwidth at 0.707 relative response) is given by

$$\textit{Bandwidth} -3\text{ dB} = f_o/Q$$

where fo is the resonant frequency and Q the circuit Q. It is also called the "half-power" bandwidth, for ease of recollection.

Parallel Resonance in Low-Q Circuits

The preceding discussion is accurate only for Qs of 10 or more. When the Q is below 10, resonance in a parallel circuit having resistance in series with the coil, as in Fig. 2-41A, is not so easily defined. There is a set of values for L and C that will make the parallel impedance a pure resistance, but with these values the impedance does not have its maximum possible value. Another set of values for L and C will make the parallel impedance a maximum, but this maximum value is not a pure resistance. Either condition could be called "resonance," so with low-Q circuits it is necessary to distinguish between maximum impedance and resistive impedance parallel resonance. The difference between these L and C values and the equal reactances of a series-resonant circuit is appreciable when the Q is in the vicinity of 5, and becomes more marked with still lower Q values.

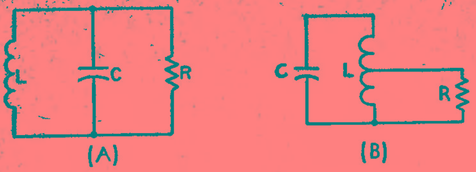


Fig. 2-43 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor *R* represents the load resistance. At B the load is tapped across part of *L*, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

Q of Loaded Circuits

In many applications of resonant circuits the only power lost is that dissipated in the resistance of the circuit itself. At frequencies below 30 MHz most of this resistance is in the coil. Within limits, increasing the number of turns in the coil increases the reactance faster than it raises the resistance, so coils for circuits in which the Q must be high are made with relatively large inductance for the frequency.

However, when the circuit delivers energy to a load (as in the case of the resonant circuits used in transmitters) the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit is shown in Fig. 2-43A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is at least ten times as great as the power lost in the inductor and capacitor, the parallel impedance of the resonant circuit itself will be so high compared with the resistance of the load that for all practical purposes the impedance of the combined circuit is equal to the load resistance. Under these conditions the Q of a parallel resonant circuit loaded by a resistive impedance is

$$Q = \frac{R}{X}$$

where R = Parallel load resistance (ohms)
X = Reactance (ohms)

Example: A resistive load of 3000 ohms is connected across a resonant circuit in which the inductive and capacitive reactances are each 250 ohms. The circuit Q is then

$$Q = \frac{R}{X} = \frac{3000}{250} = 12$$

The "effective" Q of a circuit loaded by a parallel resistance becomes higher when the reactances are decreased. A circuit loaded with a relatively low resistance (a few thousand ohms) must have low-reactance elements (large capacitance and small inductance) to have reasonably high Q.

Impedance Transformation

An important application of the parallel-resonant circuit is as an impedance-matching device in the output circuit of a vacuum-tube rf power amplifier. As described in the chapter on vacuum tubes, there is an optimum value of load resistance for each type of tube and set of operating conditions. However, the resistance of the load to which the tube is to deliver power usually is considerably lower than the value required for proper tube operation. To transform the actual load resistance to the desired value the load may be tapped across part of the coil, as shown in Fig. 2-43B. This is equivalent to connecting a higher value of load resistance across the whole circuit, and is similar in principle to impedance transformation with an iron-core transformer. In high-frequency resonant circuits the impedance ratio does not vary exactly as the square of the turns ratio, because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

When the load resistance has a very low value (say below 100 ohms) it may be connected in series in the resonant circuit (as in Fig. 2-41A, for example), in which case it is transformed to an equivalent parallel impedance as previously described. If the Q is at least 10, the equivalent parallel impedance is

$$Z_r = \frac{X^2}{r}$$

where Z_r = Resistive parallel impedance at resonance
X = Reactance (in ohms) of either the coil or capacitor
r = Load resistance inserted in series

If the Q is lower than 10 the reactance will have to be adjusted somewhat, for the reasons given in the discussion of low-Q circuits, to obtain a resistive impedance of the desired value.

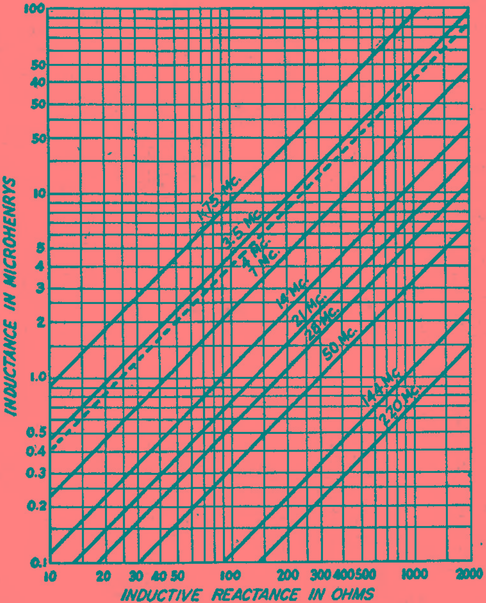


Fig. 2-44 — Reactance chart for inductance values commonly used in amateur bands from 1.75 to 220 MHz.

Reactance Values

The charts of Figs. 2-44 and 2-45 show reactance values of inductances and capacitances in the range commonly used in rf tuned circuits for the amateur bands. With the exception of the 3.5-4 MHz band, limiting values for which are shown on the charts, the change in reactance over a band, for either inductors or capacitors, is small enough so that a single curve gives the reactance with sufficient accuracy for most practical purposes.

L/C Ratio

The formula for resonant frequency of a circuit shows that the same frequency always will be obtained so long as the *product* of *L* and *C* is constant. Within this limitation, it is evident that *L* can be large and *C* small, *L* small and *C* large, etc. The relation between the two for a fixed frequency is called the L/C ratio. A high-C circuit is one that has more capacitance than "normal" for the frequency; a low-C circuit is one that has less than normal capacitance. These terms depend to a considerable extent upon the particular application considered, and have no exact numerical meaning.

LC Constants

It is frequently convenient to use the numerical value of the LC constant with a number of calculations have to be made involving different *L/C* ratios for the same frequency. The constant for any frequency is given by the following equation:

$$LC = \frac{25.330}{f^2}$$

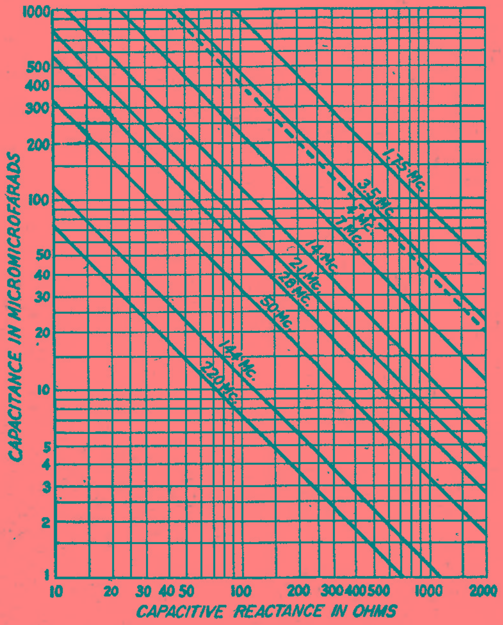


Fig. 2-45 — Reactance chart for capacitance values commonly used in amateur bands from 1.75 to 220 MHz.

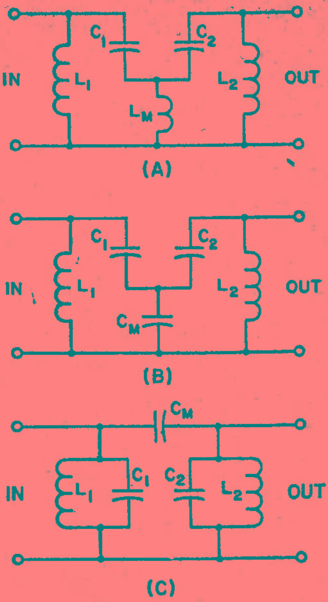


Fig. 2-46 — Three methods of circuit coupling.

where L = Inductance in microhenrys (μH)
C = Capacitance in picofarads (pF)
f = Frequency in megahertz

Example: Find the inductance required to resonate at 3650 kHz (3.65 MHz) with capacitances of 25, 50, 100 and 500 pF. The *LC* constant is

$$LC = \frac{25,330}{(3.65)2} = \frac{25,330}{13.35} = 1900$$

With 25 pF $L = 1900/C = 1900/25 = 76\ \mu H$
50 pF $L = 1900/C = 1900/50 = 38\ \mu H$
100 pF $L = 1900/C = 1900/100 = 19\ \mu H$
500 pF $L = 1900/C = 1900/500 = 3.8\ \mu H$

COUPLED CIRCUITS

Energy Transfer and Loading

Two circuits are coupled when energy can be transferred from one to the other. The circuit delivering power is called the primary circuit; the one receiving power is called the secondary circuit. The power may be practically all dissipated in the secondary circuit itself (this is usually the case in receiver circuits) or the secondary may simply act as a medium through which the power is transferred to a load. In the latter case, the coupled circuits may act as a radio-frequency impedance-matching device. The matching can be accomplished by adjusting the loading on the secondary and by varying the amount of coupling between the primary and secondary.

Coupling by a Common Circuit Element

One method of coupling between two resonant circuits is through a circuit element common to both. The three common variations of this type of coupling are shown in Fig. 2-46; the circuit element common to both circuits carries the subscript *M*. At A and B current circulating in

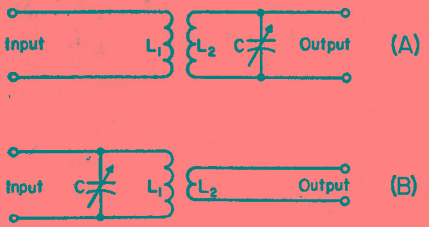


Fig. 2-47 — Single-tuned inductively coupled circuits.

L1C1 flows through the common element, and the voltage developed across this element causes current to flow in *L2C2*. At *C, CM* and *C2* form a capacitive voltage divider across *L1C1*, and some of the voltage developed across *L1C1* is applied across *L2C2*.

If both circuits are resonant to the same frequency, as is usually the case, the value of coupling reactance required for maximum energy transfer can be approximated by the following, based on $L1 = L2$, $C1 = C2$ and $Q1 = Q2$:

$$\text{(A) } L_M \approx L1/Q1; \quad \text{(B) } C_M \approx Q1C1;$$
$$\text{(C) } C_M \approx C1/Q1$$

The coupling can be increased by increasing the above coupling elements in A and C and decreasing the value in B. When the coupling is increased, the resultant bandwidth of the combination is increased, and this principle is sometimes applied to "broad-band" the circuits in a transmitter or receiver. When the coupling elements in A and C are decreased, or when the coupling element in B is increased, the coupling between the circuits is decreased below the *critical coupling* value on which the above approximations are based. Less than critical coupling will decrease the bandwidth and the energy transfer; the principle is often used in receivers to improve the selectivity.

Inductive Coupling

Figs. 2-47 and 2-48 show inductive coupling, or coupling by means of the mutual inductance between two coils. Circuits of this type resemble the iron-core transformer, but because only a part of the magnetic flux lines set up by one coil cut the turns of the other coil, the simple relationships between turns ratio, voltage ratio and impedance ratio in the iron-core transformer do not hold.

Two types of inductively-coupled circuits are shown in Fig. 2-47. Only one circuit is resonant. The circuit at A is frequently used in receivers for coupling between amplifier tubes when the tuning of the circuit must be varied to respond to signals of different frequencies. Circuit B is used principally in transmitters, for coupling a radio-frequency amplifier to a resistive load.

In these circuits the coupling between the primary and secondary coils usually is "tight" — that is, the coefficient of coupling between the coils is large. With very tight coupling either circuit operates nearly as though the device to which the untuned coil is connected were simply tapped across a corresponding number of turns on the tuned-circuit coil, thus either circuit is approximately equivalent to Fig. 2-43B.

By proper choice of the number of turns on the untuned coil, and by adjustment of the coupling, the parallel impedance of the tuned circuit may be adjusted to the value required for the proper operation of the device to which it is connected. In any case, the maximum energy transfer possible for a given coefficient of coupling is obtained when the reactance of the untuned coil is equal to the resistance of its load.

The *Q* and parallel impedance of the tuned circuit are reduced by coupling through an untuned coil in much the same way as by the tapping arrangement shown in Fig. 2-43B.

Coupled Resonant Circuits

When the primary and secondary circuits are both tuned, as in Fig. 2-48, the resonance effects in both circuits make the operation somewhat more complicated than in the simpler circuits just considered. Imagine first that the two circuits are not coupled and that each is independently tuned to the resonant frequency. The impedance of each will be purely resistive. If the primary circuit is connected to a source of rf energy of the resonant frequency and the secondary is then loosely coupled to the primary, a current will flow in the secondary circuit. In flowing through the resistance of the secondary circuit and any load that may be connected to it, the current causes a power loss. This power must come from the energy source through the primary circuit, and manifests itself in the primary as an increase in the equivalent resistance in series with the primary coil. Hence the *Q* and parallel impedance of the primary circuit are decreased by the coupled secondary. As the coupling is made greater (without changing the tuning of either circuit) the coupled resistance becomes larger and the parallel impedance of the primary continues to decrease. Also, as the coupling is made tighter the amount of power transferred from the primary to the secondary will

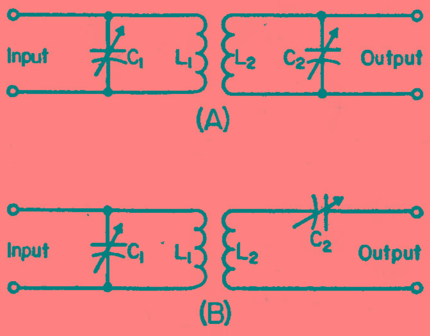


Fig. 2-48 — Inductively-coupled resonant circuits. Circuit A is used for high-resistance loads (load resistance much higher than the reactance of either L2 or C2 at the resonant frequency). Circuit B is suitable for low resistance loads (load resistance much lower than the reactance of either L2 or C2 at the resonant frequency).

increase to a maximum of one value of coupling, called critical coupling, but then decreases if the coupling is tightened still more (still without changing the tuning).

Critical coupling is a function of the Qs of the two circuits. A higher coefficient of coupling is required to reach critical coupling when the Qs are low; if the Qs are high, as in receiving applications, a coupling coefficient of a few per cent may give critical coupling.

With loaded circuits such as are used in transmitters the Q may be too low to give the desired power transfer even when the coils are coupled as tightly as the physical construction permits. In such case, increasing the Q of either circuit will be helpful, although it is generally better to increase the Q of the lower-Q circuit rather than the reverse. The Q of the parallel-tuned primary (input) circuit can be increased by decreasing the L/C ratio because, as shown in connection with Fig. 2-43, this circuit is in effect loaded by a parallel resistance (effect of coupled-in resistance). In the parallel-tuned secondary circuit, Fig. 2-48A, the Q may be increased by *increasing* the L/C ratio. There will generally be no difficulty in securing sufficient coupling, with practicable coils, if the product of the Qs of the two tuned circuits is 10 or more. A smaller product will suffice if the coil construction permits tight coupling.

Selectivity

In Fig. 2-47 only one circuit is tuned and the selectivity curve will be essentially that of a single resonant circuit. As stated, the effective Q depends upon the resistance connected to the untuned coil.

In Fig. 2-48, the selectivity is increased. It approaches that of a single tuned circuit having a Q equalling the sum of the individual circuit Qs – *if* the coupling is well below critical (this is not the condition for optimum power transfer discussed immediately above) and both circuits are tuned to resonance. The Qs of the individual circuits are affected by the degree of coupling, because each couples resistance into the other; the tighter the coupling, the lower the individual Qs and therefore the lower the over-all selectivity.

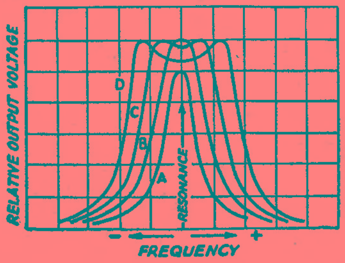


Fig. 2-49 – Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output voltage is measured across the secondary.

If both circuits are independently tuned to resonance, the over-all selectivity will vary about as shown in Fig. 2-49 as the coupling is varied. With loose coupling, A, the output voltage (across the secondary circuit) is small and the selectivity is high. As the coupling is increased the secondary voltage also increases until critical coupling, B, is reached. At this point the output voltage at the resonant frequency is maximum but the selectivity is lower than with looser coupling. At still tighter coupling, C, the output voltage at the resonant frequency decreases, but as the frequency is varied either side of resonance it is found that there are two "humps" to the curve, one on either side of resonance. With very tight coupling, D, there is a further decrease in the output voltage at resonance and the "humps" are farther away from the resonant frequency. Curves such as those at C and D are called flat-topped because the output voltage does not change much over an appreciable band of frequencies.

Note that the off-resonance humps have the same maximum value as the resonant output voltage at critical coupling. These humps are caused by the fact that at frequencies off resonance the secondary circuit is reactive and couples reactance as well as resistance into the primary. The coupled resistance decreases off resonance, and each hump represents a new condition of critical coupling at a frequency to which the primary is tuned by the additional coupled-in reactance from the secondary.

Fig. 2-50 shows the response curves for various degrees of coupling between two circuits tuned to a frequency f_o. Equal Qs are assumed in both circuits, although the curves are representative if the Qs differ by ratios up to 1.5 or even 2 to 1. In these cases, a value of $Q = \sqrt{Q1\,Q2}$ should be used.

Band-Pass Coupling

Over-coupled resonant circuits are useful where substantially uniform output is desired over a continuous band of frequencies, without readjustment of tuning. The width of the flat top of the resonance curve depends on the Qs of the two circuits as well as the tightness of coupling; the frequency separation between the humps will increase, and the curve become more flat-topped, as the Qs are lowered.

Band-pass operation also is secured by tuning the two circuits to slightly different frequencies, which gives a double-humped resonance curve even with loose coupling. This is called **stagger tuning.** To secure adequate power transfer over the frequency band it is usually necessary to use tight coupling and experimentally adjust the circuits for the desired performance.

Link Coupling

A modification of inductive coupling, called **link coupling**, is shown in Fig. 2-51. This gives the effect of inductive coupling between two coils that have no mutual inductance; the link is simply a means for providing the mutual inductance. The total mutual inductance between two coils coupled

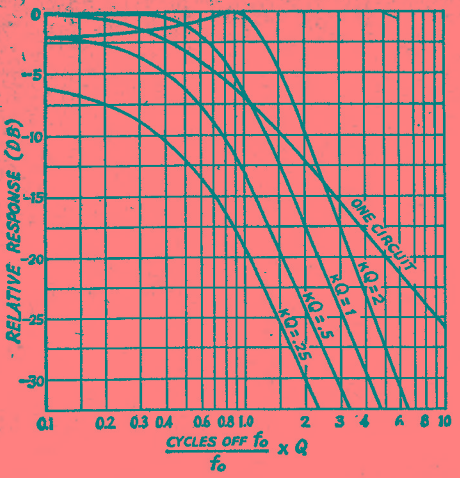


Fig. 2-50 – Relative response for a single tuned circuit end for coupled circuits. For inductively-coupled circuits (Figs. 2-46A and 2-48A),

$$k = \frac{M}{\sqrt{L1L2}}$$

where M is the mutual inductance. For capacitance-coupled circuits (Figs. 2-468 and 2-46C),

$$k \cong \frac{\sqrt{C1C2}}{C_M} \text{ end } k \cong \frac{C_M}{\sqrt{C1C2}}$$

respectively.

by a link cannot be made as great as if the coils themselves were coupled. This is because the coefficient of coupling between air-core coils is considerably less than 1, and since there are two coupling points the over-all coupling coefficient is less than for any *pair* of coils. In practice this need not be disadvantageous because the power transfer can be made great enough by making the tuned circuits sufficiently high-Q. Link coupling is convenient when ordinary inductive coupling would be impracticable for constructional reasons.

The link coils usually have a small number of turns compared with the resonant-circuit coils. The number of turns is not greatly important, because the coefficient of coupling is relatively independent of the number of turns on either coil; it is more important that both link coils should have about the *same* inductance. The length of the link between the coils is not critical if it is very small compared with the wavelength, but if the length is more than about one-twentieth of a wavelength the link operates more as a transmission line than as a means for providing mutual inductance. In such case it should be treated by the methods described in the chapter on Transmission Lines.

Fig. 2-51 – Link coupling. The mutual inductances at both ends of the link are equivelent to mutual inductance between the tuned circuits, and serve the same purpose.

IMPEDANCE-MATCHING CIRCUITS

Various combinations of L and C can be used to transform one impedance level to another and provide desirable selectivity to unwanted energy at the same time. While the simpler matching circuits use fewer components and are relatively easy to design, they lack the flexibility that is possible with more sophisticated networks.

The L network shown in Fig. 2-52 is the simplest possible impedance-matching circuit. It closely resembles an ordinary resonant circuit with the load resistance, R, either in series or parallel. The arrangement shown in Fig. 2-52A is used when the desired impedance, R_{in}, is larger than the actual load resistance, R while Fig. 2-52B is used in the opposite case. The design equations for each case are given in the figure, in terms of the circuit reactances. The reactances may be converted to inductance and capacitance by means of the formulas previously given or taken directly from the charts of Figs. 2-44 and 2-45.

The Q of an L network is found in the same way as for simple resonant circuits. That is, it is equal to XL/R or R_{IN}/XC in Fig. 2-52A, and to XL/R_{IN} or R/XC in Fig. 2-52B. The value of Q is determined by the ratio of the impedances to be matched, and cannot be selected independently. In the equations of Fig. 2-52 it is assumed that both R and R_{in} are pure resistances.

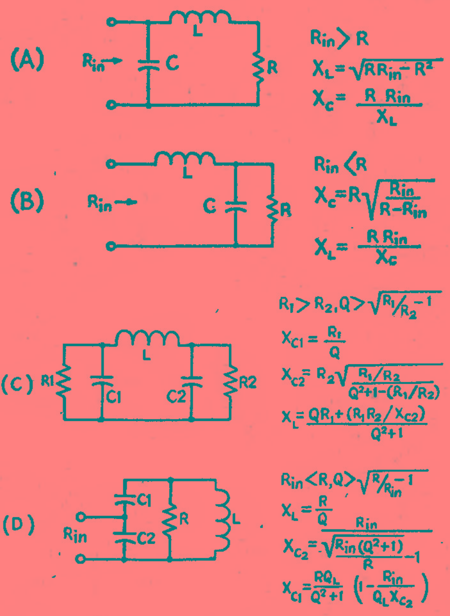


Fig. 2-52 – Impedance-matching networks adaptable to amateur work. (A) L network for matching to a lower value of resistance. (B) L network for matching to a higher resistance value. (C) pi network. (D) Versatile circuit often used in transistor- and antenna-matching networks.

The pi network shown in Fig. 2-52C is often used in the final stage of a transmitter. Different values of *L* are switched in for the appropriate band of frequencies while C1 and C2 are usually continuously variable.

In its principal application as a "tank" circuit matching a transmission line to a power amplifier tube, the load *R2* will generally have a fairly low value of resistance (up to a few hundred ohms) while *R1*, the required load for the tube, will be of the order of a few thousand ohms.

Graphical solutions for practical cases are given in the chapter on transmitter design in the discussion of plate tank circuits. The *L* and *C* values may be calculated from the reactances or read from the charts of Figs. 2-44 and 2-45.

While the pi network can be used to match a high resistance to a low one, the circuit shown in Fig. 2-52D has some attractive features. With C1 and C2 ganged and *L* variable, it is often used in matching an antenna to a transmitter (Transmatch). The inductor can be tapped to provide impedance transformation between resistances that are low in value, but nearly equal. This is often the case with many transistor circuits.

Example: Find a circuit that will match an antenna with a resistance of 1500 ohms, to a transmitter with a resistance of 50 ohms. Using the circuit shown in Fig. 2-52D, we see that *Q* has to be greater than $\sqrt{1500/50 - 1}$ or 5.38. A *Q* of 10 will satisfy this condition and a guess is made that it will also give reasonable component values. *XL* will be 1500/10 or 150 ohms.

$$XC2 = 50\sqrt{\frac{50(101)}{1500}} - 1 \text{ or } 76.9 \text{ ohms.}$$

$$XC1 = \frac{1500(10)}{101}\left(1 - \frac{50}{10(76.9)}\right) \text{ or } 138.85 \text{ ohms.}$$

If the frequency of operation was 3.7 MHz, the component values would be: $L = 6.4$ μH, C1 = 309 pF, and C2 = 559.3 pF. The guess was good since C2 is becoming large and higher values of *Q* would make this situation worse.

Quite often the load and source have reactive components along with resistance but in many instances the matching networks just discussed can still be used. The effect of these reactive components can be compensated for by changing one of the reactive elements in the matching network. For instance, if some capacitive reactance was shunted across the 1500 ohms in the last example, *L* would have to be decreased to cancel it.

FILTERS

A filter is an electrical circuit configuration (network) designed to have specific characteristics with respect to the transmission or attenuation of various frequencies that may be applied to it. There are four general types of filters: low-pass, high-pass, band-reject, and band-pass.

A **low-pass filter** is one that will permit all frequencies below a specified one, called the **cut-off frequency**, to be transmitted with little or no loss, but that will attenuate all frequencies above the **cut-off frequency**.

A **high-pass filter** similarly has a cut-off frequency, above which there is little or no loss in transmission, but below which there is considerable attenuation. Its behavior is the opposite of that of the low-pass filter.

A **band-pass filter** is one that will transmit a selected band of frequencies with substantially no loss, but that will attenuate all frequencies either higher or lower than the desired band.

A **band-reject filter** attenuates a selected band of frequencies, but allows others to be transmitted. The types that amateurs frequently encounter are commonly called traps.

The **pass band** of a filter is the frequency spectrum that is transmitted with little or no loss. The transmission characteristic is not necessarily perfectly uniform in the pass band, but the variations usually are small.

The **stop band** is the frequency region in which attenuation is desired. The attenuation may vary in the stop band, and in a simple filter usually is least near the cut-off frequency, rising to high values at frequencies considerably removed from the cut-off frequency.

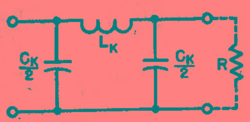
Filters are designed for a specific value of purely resistive impedance (the **terminating impedance** of the filter). When such an impedance is connected to the output terminals of the filter, the impedance looking into the input terminals has essentially the same value throughout most of the pass band. Simple filters do not give perfectly uniform performance in this respect, but the input impedance of a properly-terminated filter can be made fairly constant, as well as closer to the design value, over the pass band by using m-derived filter sections.

A discussion of filter design principles is beyond the scope of this *Handbook*, but it is not difficult to build satisfactory filters from the circuits and formulas given in Fig. 2-53. Filter circuits are built up from elementary sections as shown in the figure. These sections can be used alone or, if greater attenuation and sharper cut-off (that is, more rapid rate of rise of attenuation with frequency beyond the cut-off frequency) are required, several sections can be connected in series. In the low- and high-pass filters, *fc* represents the cut-off frequency, the highest (for the low-pass) or the lowest (for the high-pass) frequency transmitted without attenuation. In the band-pass filter designs, *f*1 is the low-frequency cut-off and *f*2 the high-frequency cut-off. The units for *L, C, R* and *f* are microhenrys, picofarads, ohms and megahertz, respectively.

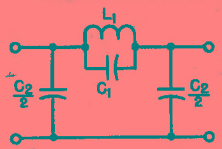
All of the types shown are "unbalanced" (one side grounded). For use in balanced circuits (e.g., 300-ohm transmission line, or push-pull audio circuits), the series reactances should be equally divided between the two legs. Thus the balanced constant-*k* π-section low-pass filter would use two inductors of a value equal to $L_k/2$, while the balanced constant-*k* π-section high-pass filter would use two capacitors each equal to $2C_k$.

If several low- (or high-) pass sections are to be used, it is advisable to use *m*-derived end sections on either side of a constant-*k* center section, although an *m*-derived center section can be used. The factor *m* determines the ratio of the cut-off

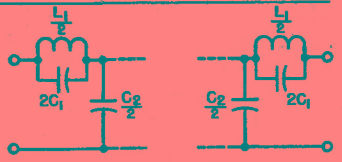
LOW-PASS FILTERS



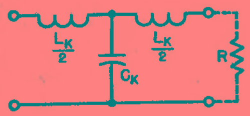
Constant-k π section



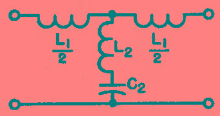
m-derived π section



m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section

m-derived end sections for use with intermediate T section

$L_K = \frac{0.318\,R}{f_c}$ $C_K = \frac{3.18(10^5)}{f_c R}$

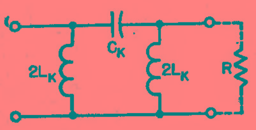
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m\,C_K$

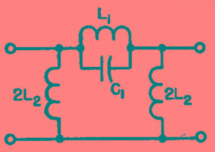
$L_1 = mL_K$ $C_1 = \frac{1-m^2}{4m} C_K$

$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m\,C_K$

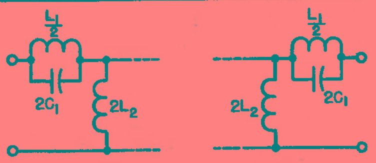
HIGH-PASS FILTERS



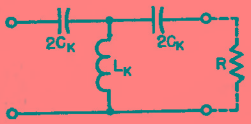
Constant-k π section



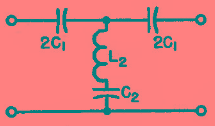
m-derived π section



m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section

m-derived end section for use with intermediate T section

$L_K = \frac{0.0796\,R}{f_c}$ $C_K = \frac{7.96(10^4)}{f_c R}$

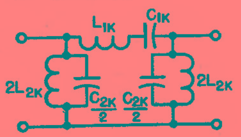
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

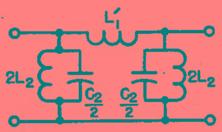
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

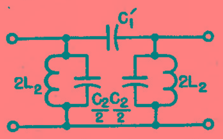
BANDPASS FILTERS



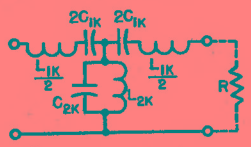
Constant-k π section



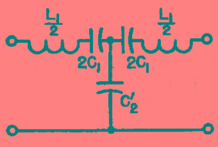
Three-element π section



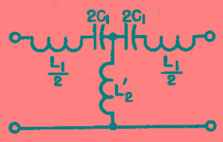
Three-element π section



Constant-k T section



Three-element T section



Three-element T section

$L_{1K} = \frac{0.318\,R}{(f_2-f_1)}$ $C_{1K} = \frac{7.96(f_2-f_1)10^4}{f_1 f_2 R}$

$L_{2K} = \frac{0.0796(f_2-f_1)R}{f_1 f_2}$ $C_{2K} = \frac{3.18(10^5)}{(f_2-f_1)R}$

$L_1 = L_{1K}$ $L_1' = \frac{0.318\,R}{f_1+f_2}$

$C_1 = \frac{7.96(f_2-f_1)10^4}{f_1 R}$ $L_2 = \frac{0.0796(f_2-f_1)R}{f_1^2}$

$C_2 = C_{2K}$ $C_2' = \frac{3.18(10^5)}{(f_1+f_2)R}$

$L_1 = \frac{0.318\,f_1 R}{f_2(f_2-f_1)}$ $C_1 = C_{1K}$

$C_1' = \frac{7.96(f_1+f_2)10^4}{f_1 f_2 R}$ $L_2 = L_{2K}$

$L_2' = \frac{0.0796(f_1+f_2)R}{f_1 f_2}$ $C_2 = \frac{3.18 f_1(10^5)}{f_2(f_2-f_1)R}$

Fig. 2-53 – Basic filter sections and design formulas. In the above formulas R is in ohms, C in farads, L in henrys, and f in cycles per second.

frequency, f_c to a frequency of high attenuation, f_∞. Where only one *m*-derived section is used, a value of 0.6 is generally used for *m*, although a deviation of 10 to 15 percent from this value is not too serious in amateur work. For a value of m = 0.6, f_∞ will be $1.25f_c$ for the low-pass filter and $0.8f_c$ for the high-pass filter. Other values can be found from

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2}$$ for the low-pass filter and

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2}$$ for the high-pass filter.

The output sides of the filters shown should be terminated in a resistance equal to *R*, and there should be little or no reactive component in the termination.

PIEZOELECTRIC CRYSTALS

A number of crystalline substances found in nature have the ability to transform mechanical strain into an electrical charge, and *vice versa*. This property is known as the piezoelectric effect. A small plate or bar cut in the proper way from a quartz crystal and placed between two conducting electrodes will be mechanically strained when the electrodes are connected to a source of voltage. Conversely, if the crystal is squeezed between two electrodes a voltage will be developed between the electrodes.

Piezoelectric crystals can be used to transform mechanical energy into electrical energy, and vice versa. They are used in microphones and phonograph pick-ups, where mechanical vibrations are transformed into alternating voltages of corresponding frequency. They are also used in headsets and loudspeakers, transforming electrical energy into mechanical vibration. Crystals of Rochelle salts are used for these purposes.

Crystal Resonators

Crystalline plates also are mechanical resonators that have natural frequencies of vibration ranging from a few thousand cycles to tens of megacycles per second. The vibration frequency depends on the kind of crystal, the way the plate is cut from the natural crystal, and on the dimensions of the plate. The thing that makes the crystal resonator valuable is that it has extremely high *Q*, ranging from a minimum of about 20,000 to as high as 1,000,000.

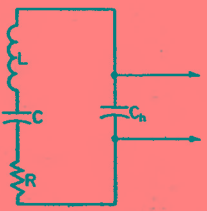


Fig. 2-54 — Equivalent circuit of a crystal resonator. *L*, *C* and *R* are the electricel equivalents of mechanical properties of the crystal; *Ch* is the capacitance of the holder plates with the crystal plate betwean them.

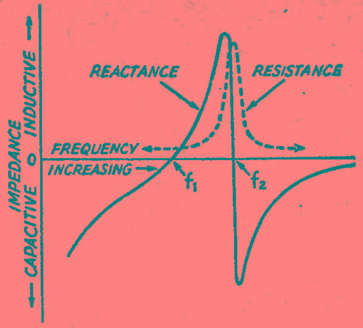


Fig. 2-55 — Reactance and resistance vs. frequency of a circuit of the type shown in Fig. 2-54. Actual values of reactance, resistance and the separation between the series- and parallel-resonant frequencies, f_1, and f_2, respectively, depend on the circuit constants.

Analogies can be drawn between various mechanical properties of the crystal and the electrical characteristics of a tuned circuit. This leads to an "equivalent circuit" for the crystal. The electrical coupling to the crystal is through the holder plates between which it is sandwiched; these plates form, with the crystal as the dielectric, a small capacitor like any other capacitor constructed of two plates with a dielectric between. The crystal itself is equivalent to a series-resonant circuit, and together with the capacitance of the holder forms the equivalent circuit shown in Fig. 2-54. At frequencies of the order of 450 kHz, where crystals are widely used as resonators, the equivalent *L* may be several henrys and the equivalent *C* only a few hundredths of a picofarad. Although the equivalent *R* is of the order of a few thousand ohms, the reactance at resonance is so high that the *Q* of the crystal likewise is high.

A circuit of the type shown in Fig. 2-54 has a series-resonant frequency, when viewed from the circuit terminals indicated by the arrowheads, determined by *L* and *C* only. At this frequency the circuit impedance is simply equal to *R*, providing the reactance of *Ch* is large compared with *R* (this is generally the case). The circuit also has a parallel-resonant frequency determined by *L* and the equivalent capacitance of *C* and *Ch* in series. Since this equivalent capacitance is smaller than *C* alone, the parallel-resonant frequency is higher than the series-resonant frequency. The separation between the two resonant frequencies depends on the ratio of *Ch* to *C*, and when this ratio is large (as in the case of a crystal resonator, where *Ch* will be a few pF, in the average case) the two frequencies will be quite close together. A separation of a kilocycle or less at 455 kHz is typical of a quartz crystal.

Fig. 2-55 shows how the resistance and reactance of such a circuit vary as the applied frequency is varied. The reactance passes through zero at both resonant frequencies, but the resistance rises to a large value at parallel resonance, just as in any tuned circuit.

Quartz crystals may be used either as simple resonators for their selective properties or as the frequency-controlling elements in oscillators as described in later chapters. The series-resonant frequency is the one principally used in the former case, while the more common forms of oscillator circuit use the parallel-resonant frequency.

PRACTICAL CIRCUIT DETAILS

COMBINED AC AND DC

Most radio circuits are built around vacuum tubes, and it is the nature of these tubes to require direct current (usually at a fairly high voltage) for their operation. They convert the direct current into an alternating current (and sometimes the reverse) at frequencies varying from well down in the audio range to well up in the super-high range. The conversion process almost invariably requires that the direct and alternating currents meet somewhere in the circuit.

In this meeting, the ac and dc are actually combined into a single current that "pulsates" (at the ac frequency) about an average value equal to the direct current. This is shown in Fig. 2-56. It is convenient to consider that the alternating current is superimposed on the direct current, so we may look upon the actual current as having two components, one dc and the other ac.

In an alternating current the positive and negative alternations have the same average amplitude, so when the wave is superimposed on a direct current the latter is alternately increased and decreased by the same amount. There is thus no *average* change in the direct current. If a dc instrument is being used to read the current, the reading will be exactly the same whether or not the ac is superimposed.

However, there is actually more power in such a combination current than there is in the direct current alone. This is because power varies as the square of the instantaneous value of the current, and when all the instantaneous squared values are averaged over a cycle the total power is greater than the dc power alone. If the ac is a sine wave having a peak value just equal to the dc, the power in the circuit is 1.5 times the dc power. An instrument whose readings are proportional to power will show such an increase.

Series and Parallel Feed

Fig. 2-57 shows in simplified form how dc and ac may be combined in a vacuum-tube circuit. In this case, it is assumed that the ac is at radio frequency, as suggested by the coil-and-capacitor tuned circuit. It is also assumed that rf current can easily flow through the dc supply; that is, the impedance of the supply at radio frequencies is so small as to be negligible.

In the circuit at the left, the tube, tuned circuit, and dc supply all are connected in series. The direct current flows through the rf coil to get to the tube; the rf current generated by the tube flows through the dc supply to get to the tuned circuit. This is series feed. It works because the impedance of the dc supply at radio frequencies is so low that it does not affect the flow of *rf* current, because the dc resistance of the coil is so low that it does not affect the flow of *direct* current.

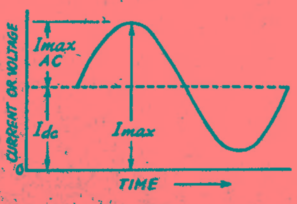


Fig. 2-56 — Pulsating dc, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

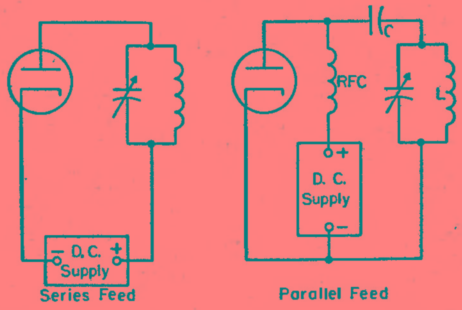


Fig. 2-57 — Illustrating series and parallel feed.

In the circuit at the right the direct current does not flow through the rf tuned circuit, but instead goes to the tube through a second coil, *RFC* (radio-frequency choke). Direct current cannot flow through *L* because a blocking capacitance, *C*, is placed in the circuit to prevent it. (Without *C*, the dc supply would be short-circuited by the low resistance of *L*.) On the other hand, the rf current generated by the tube can easily flow through *C* to the tuned circuit because the capacitance of *C* is intentionally chosen to have low reactance (compared with the impedance of the tuned circuit) at the radio frequency. The rf current cannot flow through the dc supply because the inductance of *RFC* is intentionally made so large that it has a very high reactance at the radio frequency. The resistance of *RFC*, however, is too low to have an appreciable effect on the flow of direct current. The two currents are thus in *parallel*, hence the name parallel feed.

Either type of feed may be used for both af and rf circuits. In parallel feed there is no dc voltage on the ac circuit, a desirable feature from the viewpoint of safety to the operator, because the voltages applied to tubes – particularly transmitting tubes – are dangerous. On the other hand, it is somewhat difficult to make an rf choke work well over a wide range of frequencies. Series feed is often preferred, therefore, because it is relatively easy to keep the impedance between the ac circuit and the tube low.

Bypassing

In the series-feed circuit just discussed, it was assumed that the dc supply had very low impedance at radio frequencies. This is not likely to be true in a practical power supply, partly because the normal physical separation between the supply and the rf circuit would make it necessary to use rather

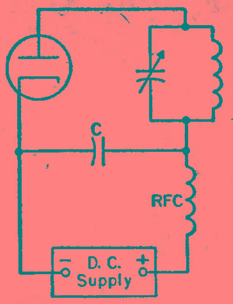


Fig. 2-58 — Typical use of a bypass capacitor and rf choke in a series-feed circuit.

long connecting wires or leads. At radio frequencies, even a few feet of wire can have fairly large reactance — too large to be considered a really "low-impedance" connection.

An actual circuit would be provided with a **bypass capacitor**, as shown in Fig. 2-58. Capacitor *C* is chosen to have low reactance at the operating frequency, and is installed right in the circuit where it can be wired to the other parts with quite short connecting wires. Hence the rf current will tend to flow through it rather than through the dc supply.

To be effective, the reactance of the bypass capacitor should not be more than one-tenth of the impedance of the bypassed part of the circuit. Very often the latter impedance is not known, in which case it is desirable to use the largest capacitance in the bypass that circumstances permit. To make doubly sure that rf current will not flow through a non-rf circuit such as a power supply, an rf choke may be connected in the lead to the latter, as shown in Fig. 2-58.

The same type of bypassing is used when audio frequencies are present in addition to rf. Because the reactance of a capacitor changes with frequency, it is readily possible to choose a capacitance that will represent a very low reactance at radio frequencies but that will have such high reactance at audio frequencies that it is practically an open circuit. A capacitance of .001 μF is practically a short circuit for rf, for example, but is almost an open circuit at audio frequencies. (The actual value of capacitance that is usable will be modified by the impedances concerned.) Capacitors also are used in audio circuits to carry the audio frequencies around a dc supply.

Distributed Capacitance and Inductance

In the discussions earlier in this chapter it was assumed that a capacitor has only capacitance and that an inductor has only inductance. Unfortunately, this is not strictly true. There is always a certain amount of inductance in a conductor, of any length, and a capacitor is bound to have a little inductance in addition to its intended capacitance. Also, there is always capacitance between two conductors or between parts of the same conductor, and thus there is appreciable capacitance between the turns of an inductance coil.

This **distributed inductance** in a capacitor and the **distributed capacitance** in an inductor have important practical effects. Actually, every capacitor is in effect a series-tuned circuit, resonant at the frequency where its capacitance and inductance have the same reactance. Similarly, every inductor is in effect a parallel tuned circuit, resonant at the frequency where its inductance and distributed capacitance have the same reactance. At frequencies well below these natural resonances, the capacitor will act like a capacitor and the coil will act like an inductor. Near the natural resonance points, the inductor will have its highest impedance and the capacitor will have its lowest impedance. At frequencies above resonance, the capacitor acts like an inductor and the inductor acts like a capacitor. Thus there is a limit to the amount of capacitance that can be used at a given frequency. There is a similar limit to the inductance that can be used. At audio frequencies, capacitances measured in microfarads and inductances measured in henrys are practicable. At low and medium radio frequencies, inductances of a few mH and capacitances of a few thousand pF are the largest practicable. At high radio frequencies, usable inductance values drop to a few μH and capacitances to a few hundred pF.

Distributed capacitance and inductance are important not only in rf tuned circuits, but in bypassing a choking as well. It will be appreciated that a bypass capacitor that actually acts like an inductance, or an rf choke that acts like a low-reactance capacitor, cannot work as it is intended they should.

Grounds

Throughout this book there are frequent references to **ground** and **ground potential**. When a connection is said to be "grounded" it does not necessarily mean that it actually goes to earth. What it means that an actual earth connection to that point in the circuit should not disturb the operation of the circuit in any way. The term also is used to indicate a "common" point in the circuit where power supplies and metallic supports (such as a metal chassis) are electrically tied together. It is general practice, for example, to "ground" the filament or heater power supplies for vacuum tubes. Since the cathode of a vacuum tube is a junction point for grid and plate voltage supplies, and since the various circuits connected to the tube elements have at least one point connected to cathode, these points also are "returned to ground." Ground is therefore a common reference point in the radio circuit. "Ground potential" means that there is no "difference of potential" — no voltage — between the circuit point and the earth.

Single-Ended and Balanced Circuits

With reference to ground, a circuit may be either **single-ended** (unbalanced) or **balanced**. In a single-ended circuit, one side of the circuit (the cold side) is connected to ground. In a balanced circuit, the electrical midpoint is connected to ground, so that the circuit has two "hot" ends each at the same voltage "above" ground.

Typical single-ended and balanced circuits are shown in Fig. 2-59. Rf circuits are shown in the

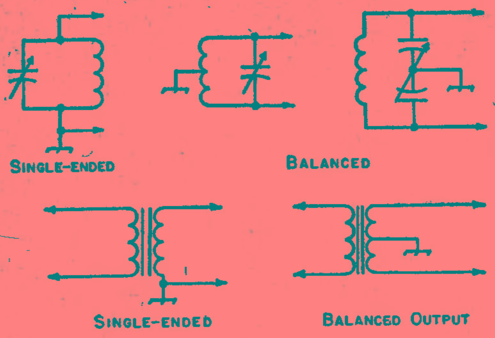


Fig. 2-59 – Single-ended and balanced circuits.

upper row, while iron-core transformers (such as are used in power-supply and audio circuits) are shown in the lower row. The rf circuits may be balanced either by connecting the center of the coil to ground or by using a "balanced" or "split-stator" capacitor and connecting its rotor to rf ground. In the iron-core transformer, one or both windings may be tapped at the center of the winding to provide the ground connection.

Shielding

Two circuits that are physically near each other usually will be coupled to each other in some degree even though no coupling is intended. The metallic parts of the two circuits form a small capacitance through which energy can be transferred by means of the electric field. Also, the magnetic field about the coil or wiring of one circuit can couple that circuit to a second through the latter's coil and wiring. In many cases these unwanted couplings must be prevented if the circuits are to work properly.

Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called shields. The electric field from the circuit components does not penetrate the shield. A metallic plate, called a baffle shield, inserted between two components also may suffice to prevent electrostatic coupling between them. It should be large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. The shielding effect for magnetic fields increases with frequency and with the conductivity and thickness of the shielding material.

A closed shield is required for good magnetic shielding; in some cases separate shields, one about each coil, may be required. The baffle shield is rather ineffective for magnetic shielding, although it will give partial shielding if placed at right angles to the axes of, and between, the coils to be shielded from each other.

Shielding a coil reduces its inductance, because part of its field is canceled by the shield. Also, there is always a small amount of resistance in the shield, and there is therefore an energy loss. This loss raises the effective resistance of the coil. The decrease in inductance and increase in resistance lower the *Q* of the coil, but the reduction in inductance and *Q* will be small if the spacing between the sides of the coil and the shield is at least half the coil diameter, and if the spacing at the ends of the coil is at least equal to the coil diameter. The higher the conductivity of the shield material, the less the effect on the inductance and *Q*. Copper is the best material, but aluminum is quite satisfactory.

For good magnetic shielding at audio frequencies it is necessary to enclose the coil in a container of high-permeability iron or steel. In this case the shield can be quite close to the coil without harming its performance.

UHF CIRCUITS

RESONANT LINES

In resonant circuits as employed at the lower frequencies it is possible to consider each of the reactance components as a separate entity. The fact that an inductor has a certain amount of self-capacitance, as well as some resistance, while a capacitor also possesses a small self-inductance, can usually be disregarded.

At the very-high and ultrahigh frequencies it is not readily possible to separate these components. Also, the connecting leads, which at lower frequencies would serve merely to join the capacitor and coil, now may have more inductance than the coil itself. The required inductance coil may be no more than a single turn of wire, yet even this single turn may have dimensions comparable to a wavelength at the operating frequency. Thus the energy in the field surrounding the "coil" may in part be radiated. At a sufficiently high frequency the loss by radiation may represent a major portion of the total energy in the circuit.

For these reasons it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 100 MHz or so.

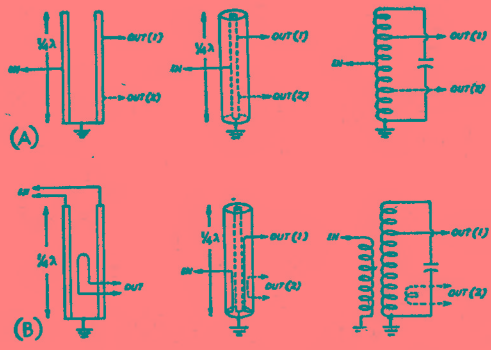


Fig. 2-60 – Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other exhibits large standing waves, as described in the section on transmission lines. When a voltage of the frequency at which such a line is resonant is applied to the open end, the response is very similar to that of a parallel resonant circuit. The equivalent relationships are shown in Fig. 2-60. At frequencies off resonance the line displays qualities comparable with the inductive and capacitive reactances of a conventional tuned circuit, so sections of transmission line can be used in much the same manner as inductors and capacitors.

To minimize radiation loss the two conductors of a parallel-conductor line should not be more than about one-tenth wavelength apart, the spacing being measured between the conductor axes. On the other hand, the spacing should not be less than about twice the conductor diameter because of "proximity effect," which causes eddy currents and an increase in loss. Above 300 MHz it is difficult to satisfy both these requirements simultaneously, and the radiation from an open line tends to become excessive, reducing the *Q*. In such case the coaxial type of line is to be preferred, since it is inherently shielded.

Representative methods for adjusting coaxial lines to resonance are shown in Fig. 2-61. At the left, a sliding shorting disk is used to reduce the effective length of the line by altering the position of the short-circuit. In the center, the same effect is accomplished by using a telescoping tube in the end of the inner conductor to vary its length and thereby the effective length of the line. At the right, two possible methods of using parallel-plate capacitors are illustrated. The arrangement with the loading capacitor at the open end of the line has the greatest tuning effect per unit of capacitance; the alternative method, which is equivalent to tapping the capacitor down on the line, has less effect on the *Q* of the circuit. Lines with capacitive "loading" of the sort illustrated will be shorter, physically, than unloaded lines resonant at the same frequency.

Two methods of tuning parallel-conductor lines are shown in Fig. 2-62. The sliding short-circuiting strap can be tightened by means of screws and nuts to make good electrical contact. The parallel-plate capacitor in the second drawing may be placed anywhere along the line, the tuning effect becoming less as the capacitor is located nearer the shorted end of the line. Although a low-capacitance variable capacitor of ordinary construction can be used, the circular-plate type shown is symmetrical and thus does not unbalance the line. It also has the further advantage that no insulating material is required.

WAVEGUIDES

A waveguide is a conducting tube through which energy is transmitted in the form of electromagnetic waves. The tube is not considered as carrying a current in the same sense that the wires of a two-conductor line do, but rather as a *boundary* which confines the waves to the enclosed space. Skin effect prevents any electromagnetic effects from being evident outside the guide. The energy is injected at one end, either through capacitive or inductive coupling or by radiation, and is received at the other end. The waveguide then merely confines the energy of the fields, which are propagated through it to the receiving end by means of reflections against its inner walls.

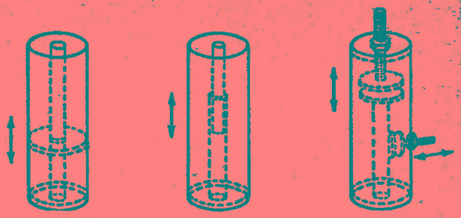


Fig. 2-61 — Methods of tuning coaxial resonant lines.

Analysis of waveguide operation is based on the assumption that the guide material is a perfect conductor of electricity. Typical distributions of electric and magnetic fields in a rectangular guide are shown in Fig. 2-63. It will be observed that the intensity of the electric field is greatest (as indicated by closer spacing of the lines of force) at the center along the *x* dimension, Fig. 2-63(B), diminishing to zero at the end walls. The latter is a necessary condition, since the existence of any electric field parallel to the walls at the surface would cause an infinite current to flow in a perfect conductor. This represents an impossible situation.

Modes of Propagation

Fig. 2-63 represents a relatively simple distribution of the electric and magnetic fields. There is in general an infinite number of ways in which the fields can arrange themselves in a guide so long as there is no upper limit to the frequency to be transmitted. Each field configuration is called a mode. All modes may be separated into two general groups. One group, designated *TM* (transverse magnetic), has the magnetic field entirely transverse to the direction of propagation, but has a component of electric field in that direction. The other type, designated *TE* (transverse electric) has the electric field entirely transverse, but has a component of magnetic field in the direction of propagation. *TM* waves are sometimes called *E* waves, and *TE* waves are sometimes called *H* waves, but the *TM* and *TE* designations are preferred.

The particular mode of transmission is identified by the group letters followed by two subscript

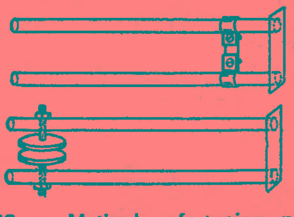


Fig. 2-62 — Methods of tuning parallel-type resonant lines.

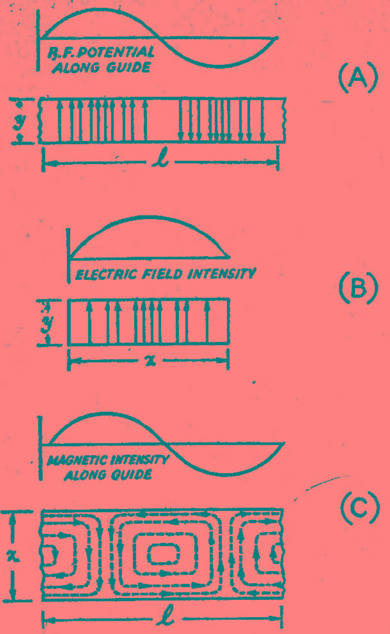


Fig. 2-63 – Field distribution in a rectangular waveguide. The $TE_{1,0}$ mode of propagation is depicted.

numerals; for example, $TE_{1,0}$, $TM_{1,1}$, etc. The number of possible modes increases with frequency for a given size of guide. There is only one possible mode (called the dominant mode) for the lowest frequency that can be transmitted. The dominant mode is the one generally used in practical work.

Waveguide Dimensions

In the rectangular guide the critical dimension is x in Fig. 2-63; this dimension must be more than one-half wavelength at the lowest frequency to be transmitted. In practice, the y dimension usually is made about equal to $1/2\,x$ to avoid the possibility of operation at other than the dominant mode.

Other cross-sectional shapes than the rectangle can be used, the most important being the circular pipe. Much the same considerations apply as in the rectangular case.

Wavelength formulas for rectangular and circular guides are given in the following table, where x is the width of a rectangular guide and r is the radius of a circular guide. All figures are in terms of the dominant mode.

	Rectangular	*Circular*
Cutoff wavelength	$2x$	$3.41r$
Longest wavelength transmitted with little attenuation	$1.6r$	$3.2r$
Shortest wavelength before next mode becomes possible	$1.1x$	$2.8r$

Cavity Resonators

Another kind of circuit particularly applicable at wavelengths of the order of centimeters is the cavity resonator, which may be looked upon as a section of a waveguide with the dimensions chosen so that waves of a given length can be maintained inside.

Typical shapes used for resonators are the cylinder, the rectangular box and the sphere, as shown in Fig. 2-64. The resonant frequency depends upon the dimensions of the cavity and the mode of oscillation of the waves (comparable to the transmission modes in a waveguide). For the lowest modes the resonant wavelengths are as follows:

Cylinder	$2.61r$
Square box	$1.41l$
Sphere	$2.28r$

The resonant wavelengths of the cylinder and square box are independent of the height when the height is less than a half wavelength. In other modes of oscillation the height must be a multiple of a half wavelength as measured inside the cavity. A cylindrical cavity can be tuned by a sliding shorting disk when operating in such a mode. Other tuning methods include placing adjustable tuning paddles or "slugs" inside the cavity so that the standing-wave pattern of the electric and magnetic fields can be varied.

A form of cavity resonator in practical use is the re-entrant cylindrical type shown in Fig. 2-65. In construction it resembles a concentric line closed at both ends with capacitive loading at the top, but the actual mode of oscillation may differ considerably from that occuring in coaxial lines. The resonant frequency of such a cavity depends upon the diameters of the two cylinders and the distance d between the cylinder ends.

Compared with ordinary resonant circuits, cavity resonators have extremely high Q. A value of Q of the order of 1000 or more is readily obtainable, and Q values of several thousand can be secured with good design and construction.

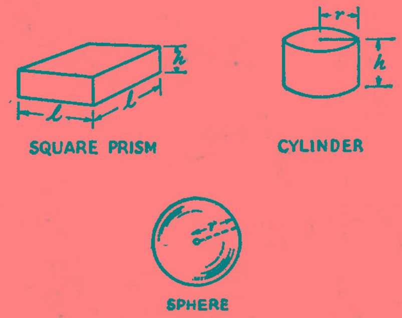


Fig. 2-64 – Forms of cavity resonators.

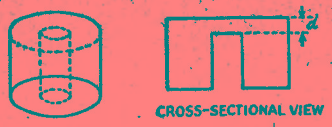


Fig. 2-65 — Re-entrant cylindrical cavity resonator.

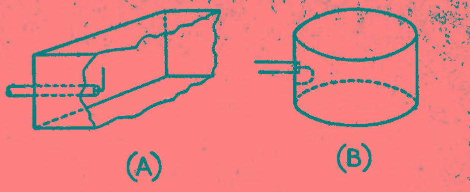


Fig. 2-66 — Coupling to waveguides and resonators.

Coupling to Waveguides and Cavity Resonators

Energy may be introduced into or abstracted from a waveguide or resonator by means of either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods of coupling to which are shown in Fig. 2-66. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

Coupling can be varied by turning the probe or loop through a 90-degree angle. When the probe is perpendicular to the electric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its minimum value.

MODULATION, HETERODYNING, AND BEATS

Since one of the most widespread uses of radio frequencies is the transmission of speech and music, it would be very convenient if the audio spectrum to be transmitted could simply be shifted up to some radio frequency, transmitted as radio waves, and shifted back down to audio at the receiving point. Suppose the audio signal to be transmitted by radio is a pure 1000-hertz tone, and we wish to transmit the signal at 1 MHz (1,000,000 hertz. One possible way to do this might be to add 1.000 MHz and 1 kHz together, thereby obtaining a radio frequency of 1.001 MHz. No simple method for doing this directly has been devised, although the *effect* is obtained and used in "single-sideband transmission."

When two different frequencies are present simultaneously in an ordinary circuit (specifically, one in which Ohm's Law holds) each behaves as though the other were not there. The total or resultant voltage (or current) in the circuit will be the sum of the instantaneous values of the two at every instant. This is because there can be only one value of current or voltage at any single point in a circuit at any instant. Figs. 2-67A and B show two such frequencies, and C shows the resultant. The amplitude of the 1-MHz current is not affected by the presence of the 1-kHz current, but the axis is shifted back and forth at the 1-kHz rate. An attempt to transmit such a combination as a radio wave would result in only the radiation of the

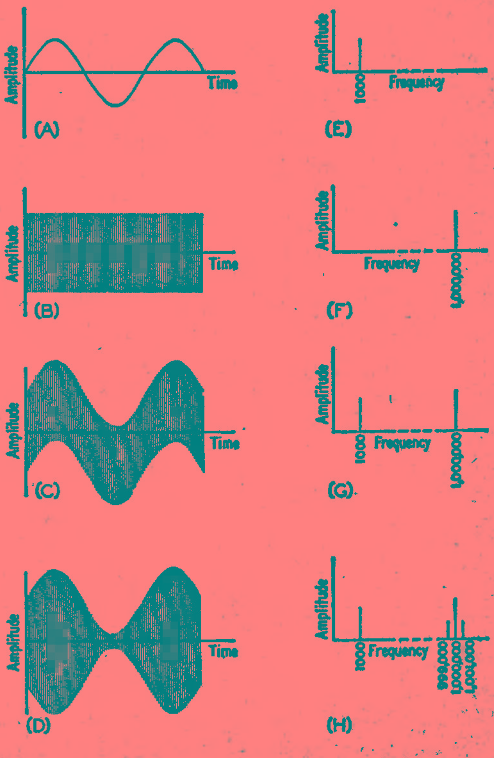


Fig. 2-67 — Amplitude-vs.-time and amplitude-vs.-frequency plots of verious signels. (A) 1-1/2 cycles of an audio signal, assumed to be 1000 hz in this example. (B) A radio-frequency signal, assumed to be 1 MHz; 1500 hertz are completed during the same time as the 1-1/2 cycles in A, so they cannot be shown accuretely. (C) The signals of A and B in the same circuit; each maintains its own identity. (D) The signals of A and B in a circuit where the amplitude of A can control the amplitude of B. The 1-MHz signal is modulated by the 1000-hz signal.

E, F, G and H show the spectrums for the signals in A, B, C end D, respectively. Note the new frequencies in H, resulting from the modulation process.

1-MHz frequency, since the 1-kHz frequency retains its identity as an audio frequency and will not radiate.

There are devices, however, which make it possible for one frequency to control the amplitude of the other. If, for example, a 1-kHz tone is used to control a 1-MHz signal, the maximum rf output will be obtained when the 1-kHz signal is at the peak of one alternation and the minimum will occur at the peak of the next alternation. The process is called **amplitude modulation**, and the effect is shown in Fig. 2-67D. The resultant signal is now entirely at radio frequency, but with its amplitude varying at the modulation rate (1 kHz). Receiving equipment adjusted to receive the 1-MHz rf signal can reproduce these changes in amplitude, and reveal what the audio signal is, through a process called **detection**.

It might be assumed that the only radio frequency present in such a signal is the original 1.000 MHz, but such is not the case. Two new frequencies have appeared. These are the sum (1.00 + .001) and the difference (1.000 – .001) of the two, and thus the radio frequencies appearing after modulation are 1.001, 1.000 and .999 MHz.

When an audio frequency is used to control the amplitude of a radio frequency, the process is generally called "amplitude modulation," as mentioned, but when a radio frequency modulates another radio frequency it is called **heterodyning**. The processes are identical. A general term for the sum and difference frequencies generated during heterodyning or amplitude modulation is **"beat frequencies,"** and a more specific one is **upper side frequency**, for the sum, and **lower side frequency** for the difference.

In the simple example, the modulating signal was assumed to be a pure tone, but the modulating signal can just as well be a *band* of frequencies making up speech or music. In this case, the side frequencies are grouped into the **upper sideband** and the **lower sideband**. Fig. 2-67H shows the side frequencies appearing as a result of the modulation process.

Amplitude modulation (a-m) is not the only possible type nor is it the only one in use. Such signal properties as phase and frequency can also be modulated. In every case the modulation process leads to the generation of a new set (or sets) of radio frequencies symmetrically disposed about the original radio (carrier) frequency.

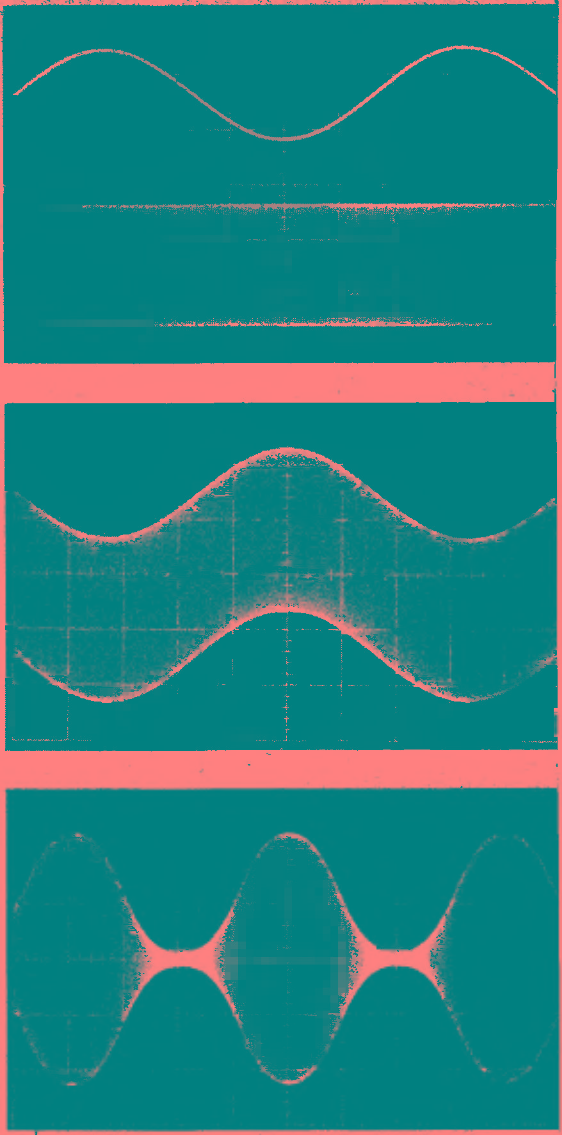


Fig. 2-68 — Actual oscilloscope photograph showing the signals described in the text and shown in the drawings of Fig. 2-67.

TOROIDAL INDUCTORS AND TRANSFORMERS

With many builders, miniaturization is the watchword. This is especially true when working with solid-state and etched-circuit projects. One of the deterrents encountered in designing small-volume equipment is the squeezing in of bulky inductors – slug-tuned or air wound – into a compact assembly. Toroids offer a practical solution to the problem of mass. The good points do not end there, however; toroidal-wound inductors not only fit into small places, they offer exceptionally high values of tuned-circuit *Q*, a definite attribute when selectivity in an important consideration in equipment performance. Ordinarily, air-wound inductors which provide comparable *Q* are many times larger than are their toroidal kinsmen. The correct type of core material must be used in order to realize the best possible *Q* at a particular frequency.

Minimum interaction between the tuned stages of a given piece of equipment is usually of paramount importance to the builder. Here is where the toroid performs well; a toroidal inductor is self-shielding. That is to say, its magnetic flux is very nearly all contained within the coil itself. This feature cuts down stray inductive coupling between adjacent circuits and permits the toroid to be mounted physically close to other components – including the chassis and cabinet walls – without impairment of its efficiency. The latter is not true of ordinary rf or af inductors. Because the flux is contained within the toroid coil, tighter coupling between windings, when a primary and secondary are used, is possible.

The high permeability of ferrite toroid cores permits the user to employ fewer turns in the tuned-circuit inductor. With fewer turns of wire required, larger wire gauges can be used, with a resultant reduction in heating and I^2R losses. This feature is especially beneficial in transistorized equipment where high collector currents are frequently required.

It is best to understand that the word "toroidal" refers to a physical format – doughnut shape – rather than to a specific device or type of material. Toroid cores come in a host of sizes, are manufactured by many firms (each with a different identifying code for the type of core material used), and are fashioned from a wide variety of materials. Some cores are made by rolling up great lengths of thin silicon steel tape (Hypersil) into a toroidal form. Such cores are held together by means of plastic covers, or are wrapped with glass tape which holds the core intact while insulating it from the wire which is wound on it. This type of core is commonly used for low-frequency power applications such as dc-to-dc, and dc-to-ac converters. For audio and rf applications powdered iron and ferrite (a newer type of ceramic) material are generally used. Ferrite acts like an insulating material, making it unnecessary in all instances to place a layer of tape between the core and the winding of the transformer or inductor.

Choosing a Core

There is no simple rule that can be used for selecting a toroid core for a particular job. Many things must be considered notably the intended frequency of operation, the operating frequency versus the physical size and permeability of the core, and whether or not the core will be used in a small- or large-signal tuned circuit. The higher the permeability rating of the material, the fewer will be the number of turns required to obtain a specific inductance value. For example: if a core of certain size has a permeability rating of 400, it might require, say, 25 turns of wire to give an inductance of 10 μH. Therefore, where minimum I^2R loss in the winding is desirable, the higher permeability is better. A core with a larger cross-sectional area (computed from inside diameter, outside diameter, and core height) will reduce the required number of turns also. These are but a few possibilites to consider when selecting a core. Q1 material is rated for rf applications up to 10 MHz, Q2 stock is good to 50 MHz, and Q3 ferrite is rated to 225 MHz. These three ranges handle most rf needs.[1] If the improper material is chosen for a given frequency of operation, the core material will not provide a high-Q inductor. In fact, the wrong material can completely ruin a tuned circuit. If too large a core (physical size) is used in the upper hf region, or at vhf, it may be impossible to wind a suitable coil on the toroid because so little wire will be required to provide the needed value of inductance. For this reason, the smaller cores, and those with low permeability ratings, should be used in the upper frequency range.

It is helpful to have some knowledge of the core types offered by the various companies before ordering a toroid for a particular project. Indiana General offers a specification sheet for each of their core materials (see Table I). Each sheet lists such data as permeability, flux density, residual magnetism, usable frequency range, and the loss factor at a specified frequency. Bulletin 101A lists the physical dimensions of their cores and also gives the cross-sectional area of each model in square inches. With this information one can calculate the required number of turns for a specific inductance value, using a selected core size. With the foregoing information at our disposal, the formula given here will enable the constructor to determine the inductance of a toroid when the number of turns is known:

$$L = \left(0.0046\,\mu N^2 h \ \log_{10} \frac{OD}{ID} \right) \mu H$$

Where L = inductance
μ = permeability of the material
N = number of turns
OD = outer diameter of core (cm.)
ID = inner diameter of core (cm.)
h = height of core (in cm.)

To obtain dimensions in centimeters, multiply inches by 2.54. The inductance nomogram given in Fig. 2-70 can be used when designing toroidal inductors which are to be wound on the standard cores offered by Indiana General.

Specific Applications

Because toroids can be used in circuits that handle anything from microwatts to kilowatts, they can be put to good use in almost any tuned-circuit or transformer application.

Most amateurs are familiar with balun transformers, having used them at one time or another in their antenna systems. Toroids find widespread use as balun transformers because they provide a broad-band transformer that is compact and offers good power-transfer efficiency. An article which describes how to contruct homemade toroidal baluns was published in August 1964 *QST*. Core size with respect to four different power levels – 150 to 1000 watts – is treated in the article.

Toroidal inductors are useful when applied to circuits in which a high degree of selectivity is desired. A high-Q toroidal tuned circuit in the rf and mixer stages of a communications receiver can aid image rejection more than is possible with conventional slug-tuned inductors.

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[1] Q1, Q2, and Q3 designations used here are those assigned to cores made by Indiana General Corp., Keasbey, NJ 08832. Other manufactures of ferramic materials use different identifying codes.

Another application for toroidal inductors is in transistorized transmitting and receiving equipment – and in some vacuum-tube circuits – where broad-band input, interstage, or output rf transformers are desired. Toroids can be used in such circuits to provide good efficiency and small physical size. The broad-band transformer requires no tuning controls when properly designed for a given frequency range – a particularly useful feature in mobile equipment. It is not difficult to design a broad-band transformer [2] that will work over a range of 3 to 30 MHz, but one must take precautions against the radiation of harmonic energy when using this kind of transformer in the final stage of a transmitter.

Compact equipment calls for the close spacing of component parts, often requiring that the tuned circuits of several stages be in close physical proximity. This sort of requirement often leads to electrical instability of one or more of the stages, because of unwanted interstage coupling, thus impairing the performance of the equipment. Because the toroidal transformer or inductor is self-shielding, it is possible to place the tuned circuits much closer together than when using conventional inductors. The self-shielding feature also makes it possible to mount a toroid against a circuit board, or against a metal chassis or cabinet wall, without significantly affecting their *Q*. Normally, the most noticeable effect of moving a toroid closer to or farther away from a metal surface is a change in overall circuit capacitance, which in turn slightly affects the resonant frequency of the toroidal tuned circuit. Because fewer turns of wire are needed for a toroid coil than for ordinary ai-wound or slug-tuned inductors, the assembly can be made extremely compact – a much sought-after feature in miniaturized equipment.

Inductors and transformers which are wound on toroid cores are subject to the same general conditions that are common to the laminated iron-core types treated earlier in this chapter. A sufficient amount of cross-sectional area is necessary for a given amount of power in order to prevent saturation and heating. Either of these conditions will seriously impair the efficiency of a circuit. When toroids are used in circuits where high pk-pk rf voltage is (or can be) present, the core should be wrapped with glass tape or some insulating material of similar characteristics. Teflon-insulated wire should be used to prevent flashover between turns, or between the winding and the core.

Additional design data and information on making one's own toroid cores are given in, "Toroidal-Wound Inductors," *QST* for January 1968, page 11. Industrial data files and application notes are available from the manufacturers of ferrite products.[3]

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[2] C.L. Ruthroff, "Some Broadband Transformers", *Proc. IREs* Vol. 47, p. 137, Aug. 1959.

[3] Indiana General Corp., Electronics Div./ Ferrites, Keasbey, NJ 08832. Also, Ferroxcube Corp. of America, Saugerties, NY 12477 and Amidon Associates, 12033 Otsego St., N. Hollywood, CA 91607.

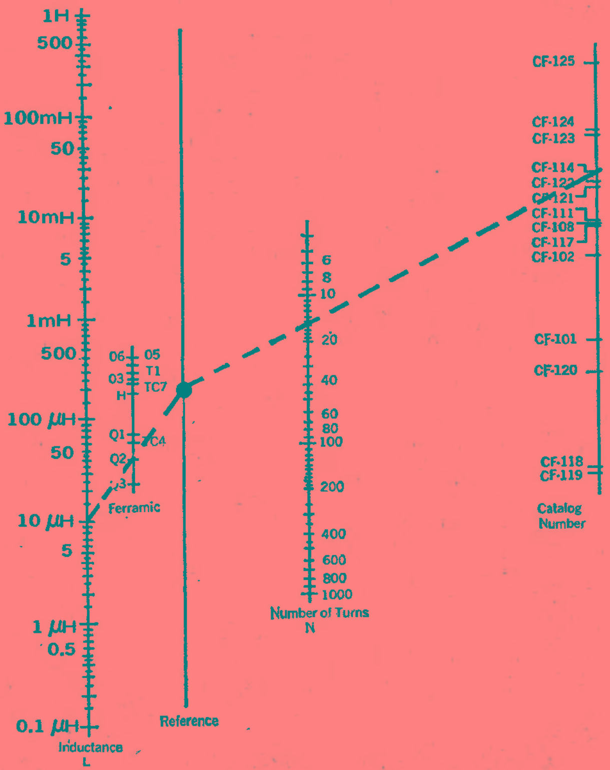

Fig. 2-69 – Nomograph which can be used to calculate the number of turns required for a specific inductance once the type of core (Indiana General) is known. Draw a line of inductance, L through the marker which indicetes the core material being used, Q1, Q2, Q3, etc. Complete this line until it intersects the Reference line. Now draw a line from the intersect point on the Raference line to the catelog number line of the nomogram (CF number of the core). This line will cross the Number of Turns (N) line, indicating the number of turns needed. Example shown 15-turn winding required for 10μH inductance on CF-114 core of Q2 material. (Nomogram courtesy of Indiana General.)

Chapter 3

Vacuum-Tube Principles

CURRENT IN A VACUUM

The outstanding difference between the vacuum tube and most other electrical devices is that the electric current does not flow through a conductor but through empty space – a vacuum. This is only possible when "free" electrons – that is, electrons that are not attached to atoms – are somehow introduced into the vacuum. Free electrons in an evacuated space will be attracted to a positively charged object within the same space, or will be repelled by a negatively charged object. The movement of the electrons under the attraction or repulsion of such charged objects constitutes the current in the vacuum.

The most practical way to introduce a sufficiently large number of electrons into the evacuated space is by **thermionic emission.**

Thermionic Emission

If a piece of metal is heated to incandescence in a vacuum, electrons near the surface are given enough energy of motion to fly off into the surrounding space. The higher the temperature, the greater the number of electrons emitted. The name for the emitting metal is **cathode.**

If the cathode is the only thing in the vacuum, most of the emitted electrons stay in its immediate vicinity, forming a "cloud" about the cathode. The reason for this is that the electrons in the space, being negative electricity, for a negative charge (space charge) in the region of the cathode. The space charge repels those electrons nearest the cathode, tending to make them fall back on it.

Transmitting tubes are in the back and center rows. Receiving tubes are in the front row (l. to r.): miniature, pencil, planar triode (two), Nuvistor and 1-inch diameter cathode-ray tube.

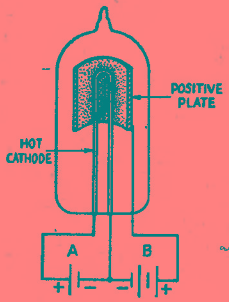

Fig. 3-1 – Conduction by thermionic emission in e vacuum tube. The A battery is used to heet the cathode to a temperature that will cause it to emit electrons. The B battery makes the plate positive with respect to the cathode, thereby causing the emitted electrons to be attracted to the plate. Electrons captured by the plate flow back through the B battery to the cathode.

Now suppose a second conductor is introduced into the vacuum, but not connected to anything else inside the tube. If this second conductor is given a positive charge by connecting a voltage source between it and the cathode, as indicated in Fig. 3-1, electrons emitted by the cathode are attracted to the positively charged conductor. An electric current then flows through the circuit formed by the cathode, the charged conductor, and the voltage source. In Fig. 3-1 this voltage source is a battery ("B" battery); a second battery ("A" battery) is also indicated for heating the cathode to the proper operating temperature.

The positively charged conductor is usually a metal plate or cylinder (surrounding the cathode) and is called an **anode** or **plate.** Like the other working parts of a tube, it is a **tube element** or **electrode.** The tube shown in Fig. 3-1 is a **two-element** or **two-electrode** tube, one element being the cathode and the other the anode or plate.

Since electrons are negative electricity, they will be attracted to the plate *only* when the plate is positive with respect to the cathode. If the plate is given a negative charge, the electrons will be repelled back to the cathode and no current will flow. The vacuum tube therefore can conduct *only in one direction.*

Cathodes

Before electron emission can occur, the cathode must be heated to a high temperature. However, it is not essential that the heating current flow

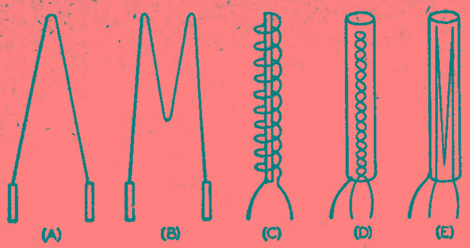

Fig. 3-2 — Types of cathode construction. Directly heated cathodes or "filaments" are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting tube type. The indirectly heated cathodes at D and E show two types of heater construction, one a twisted loop and the other bunched heater wires. Both types tend to cancel the magnetic fields set up by the current through the heater.

through the actual material that does the emitting; the filament or heater can be electrically separate from the emitting cathode. Such a cathode is called **indirectly heated**, while an emitting filament is called a **directly heated** cathode. Fig. 3-2 shows both types in the forms which they commonly take.

Much greater electron emission can be obtained, at relatively low temperatures, by using special cathode materials rather than pure metals. One of these is **thoriated tungsten** or tungsten in which thorium is dissolved. Still greater efficiency is achieved in the oxide-coated cathode, a cathode in which rare-earth oxides form a coating over a metal base.

Although the oxide-coated cathode has the highest efficiency, it can be used successfully only in tubes that operate at rather low plate voltages. Its use is therefore confined to receiving-type tubes and to the smaller varieties of transmitting tubes. The thoriated filament, on the other hand, will operate well in high-voltage tubes.

Plate Current

If there is only a small positive voltage on the plate, the number of electrons reaching it will be small because the space charge (which is negative) prevents those electrons nearest the cathode from being attracted to the plate. As the plate voltage is increased, the effect of the space charge is increasingly overcome and the number of electrons attracted to the plate becomes larger. That is, the **plate current increases with increasing plate voltage.**

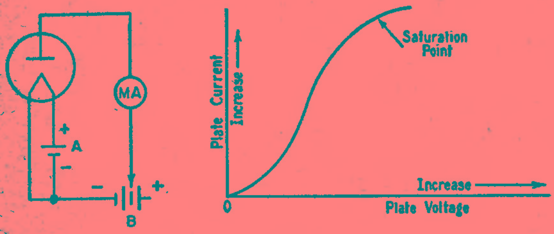

Fig. 3-3 — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

Fig. 3-3 shows a typical plot of plate current *vs.* plate voltage for a two-element tube or diode. A curve of this type can be obtained with the circuit shown, if the plate voltage is increased in small steps and a current reading taken (by means of the current-indicating instrument — a milliammeter) at each voltage. The plate current is zero with no plate voltage and the curve rises until a **saturation point** is reached. This is where the positive charge on the plate has substantially overcome the space charge and almost all the electrons are going to the plate. At higher voltages the plate current stays at practically the same value.

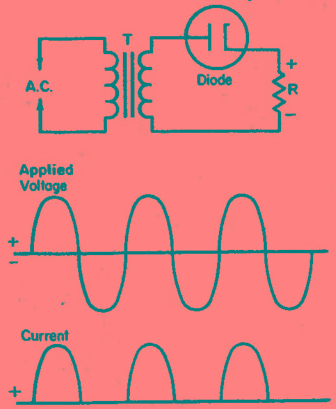

Fig. 3-4 — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.

The plate voltage multiplied by the plate current is the power input to the tube. In a circuit like that of Fig. 3-3 this power is all used in heating the plate. If the power input is large, the plate temperature may rise to a very high value (the plate may become red or even white hot). The heat developed in the plate is radiated to the bulb of the tube, and in turn radiated by the bulb to the surrounding air.

RECTIFICATION

Since current can flow through a tube in only one direction, a diode can be used to change alternating current into direct current. It does this by permitting current to flow only when the anode is positive with respect to the cathode. There is no current flow when the plate is negative.

Fig. 3-4 shows a representative circuit. Alternating voltage from the secondary of the transformer, *T*, is applied to the diode tube in series with a load resistor, *R*. The voltage varies as is usual with ac, but current flows through the tube and *R* only when the plate is positive with respect to the cathode — that is, during the half-cycle when the upper end of the transformer winding is positive. During the negative half-cycle there is simply a gap in the current flow. This rectified alternating current therefore is an *intermittent* direct current.

The load resistor, *R*, represents the actual circuit in which the rectified alternating current does work. All tubes work with a load of one type or another; in this respect a tube is much like a generator or transformer. A circuit that did not provide a load for the tube would be like a short-circuit across a transformer; no useful purpose would be accomplished and the only result would be the generation of heat in the transformer. So it is with vacuum tubes; they must cause power to be developed in a load in order to serve a useful purpose. Also, to be *efficient* most of the power must do useful work in the load and not be used in heating the plate of the tube. Thus the voltage drop across the load should be much higher than the drop across the diode.

With the diode connected as shown in Fig. 3-4, the polarity of the current through the load is as indicated. If the diode were reversed, the polarity of the voltage developed across the load *R* would be reversed.

VACUUM-TUBE AMPLIFIERS

TRIODES

Grid Control

If a third element – called the control grid, or simply grid – is inserted between the cathode and plate as in Fig. 3-5, it can be used to control the effect of the space charge. If the grid is given a positive voltage with respect to the cathode, the positive charge will tend to neutralize the negative space charge. The result is that, at any selected plate voltage, more electrons will flow to the plate than if the grid were not present. On the other hand, if the grid is made negative with respect to the cathode the negative charge on the grid will add to the space charge. This will reduce the number of electrons that can reach the plate at any selected plate voltage.

The grid is inserted in the tube to control the space charge and not to attract electrons to itself, so it is made in the form of a wire mesh or spiral. Electrons then can go through the open spaces in the grid to reach the plate.

Characteristic Curves

For any particular tube, the effect of the grid voltage on the plate current can be shown by a set of characteristic curves. A typical set of curves is shown in Fig. 3-6, together with the circuit that is used for getting them. For each value of plate voltage, there is a value of negative grid voltage that will reduce the plate current to zero; that is, there is a value of negative grid voltage that will cut off the plate current.

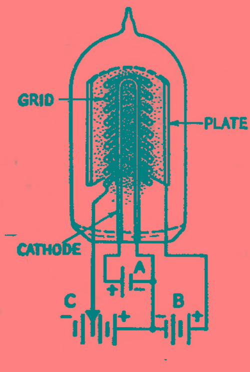

Fig. 3-5 – Construction of an elementary triode vacuum tube, showing the directly-heated cathode (filament), grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.

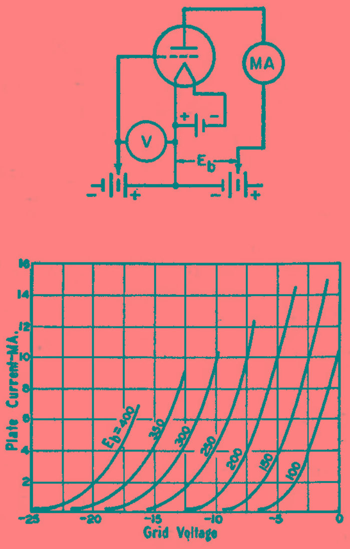

Fig. 3-6 – Grid-voltage-vs.-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

The curves could be extended by making the grid voltage positive as well as negative. When the grid is negative, it repels electrons and therefore none of them reaches it; in other words, no current flows in the grid circuit. However, when the grid is positive, it attracts electrons and a current (grid current) flows, just as current flows to the positive plate. Whenever there is grid current there is an accompanying power loss in the grid circuit, but so long as the grid is negative no power is used.

It is obvious that the grid can act as a valve to control the flow of plate current. Actually, the grid has a much greater effect on plate current flow than does the plate voltage. A small change in grid voltage is just as effective in bringing about a given change in plate current as is a large change in plate voltage.

The fact that a small voltage acting on the grid is equivalent to a large voltage acting on the plate indicates the possibility of amplification with the