ELECTRONIC TECHNOLOGY SERIES

TRANSFORMERS

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TRANSFORMERS

edited by

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PREFACE

The ease with which transformers transfer energy from one circuit to another by electromagnetic induction allows their common usage in all aspects of electrical and electronic technology. The absence of moving parts, the transfer of energy without change of frequency, high efficiency, and the relatively little care required for maintenance because of a transformer's simple and durable construction, indicate increasing roles for these devices in communications and power work. Thus, it is necessary for those working with electronics to understand the essential relationships of transformer theory.

This book discusses and evaluates transformer theory and the varied types of transformer operations and applications. Transformer theory is presented with simple mathematical treatment, which permits sufficiently extensive analysis and allows the interested technician or student to develop full comprehension. Adequate information is given relating to broad concepts and information designed for ready use. Detailed descriptions of a small number of selected major topics are presented and, through presentation of practical situations, equipment, and problems, the reader is afforded an opportunity to apply the principles he has learned.

Specific attention is given to transformer magnetics, flux density, magnetic intensity, a review of magnetic units, permeability of free space, and relative permeability. Magnetization curves, hysteresis loop, core losses and their measurement are also discussed. In addition, current and voltage waveforms in transformer primaries, transformer shielding, design, and construction are analyzed. The theory of basic operation efficiency, coil currents, coupling and mutual inductance, leakage inductance, and distributed capacitance receive particular attention. Transformer size and efficiency, core materials, power transformers, pertinencies of filter systems, filament transformer ratings, transformer primary control, distribution systems are evaluated. Special emphasis is placed on voltage and current relationships in three-phase systems, audio transformer types, equivalent circuits, the impedence ratio of audio transformers, parallel-feed coupling method, and considerations of high-frequency transformers. The varied uses of undercoupling, critical coupling, transitional coupling, and overcoupling are dwelt upon. Gain-bandwidth factors, special transformers and their applications, saturable reactors, self-saturating saturable reactors, voltage-regulating transformers, and balancing transformers are also discussed. Special emphasis is devoted to problems relating to the pertinent areas of transformer theory.

Mastery of the material presented herein will insure the student an adequate basis to undertake work requiring a pre-requisite knowledge of transformers.

Grateful acknowledgment is made to the staff of New York Institute of Technology for its assistance in the preparation of the manuscript of this book.

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Chapter 1

TRANSFORMER MAGNETICS

1. Introduction

The word *transformer* has different meanings in different fields. For example, one college physics text defines a transformer as "a device used to change an alternating potential difference from one value to another." An industrial electronics book states that "any two coils arranged in a way such that they have mutual inductance with respect to each other can be called a transformer." A recent secondary school book on electricity refers to a transformer as a device "having a primary and secondary coil with a common laminated iron core." Finally, a radio textbook defines a transformer as a device in which "two or more coils are arranged so that energy may be transferred from one circuit to another by electromagnetic induction."

Of these definitions, the last is the most general and inclusive. Accepting this definition, it is logical to divide transformers into types, each classification based upon application. The radio engineer thinks of transformers in terms of air-core or powdered-iron types. He often intentionally places two coils so that only a small fraction of the magnetic field produced by one coil passes through the other. These coils are *loosely-coupled.* The power engineer is concerned only with iron-core transformers, since his interest lies in obtaining a very high degree of coupling. He would like to have all the magnetic lines produced by the current in one coil linked with the second coil. On the other hand, an engineer interested in general electronics deals with all types of transformers —from loosely-coupled air-core types, through the various degrees of coupling found in intermediate-frequency (i-f) transformers, to power transformers having almost 100% linkage.

Theoretically, the degree of coupling or the fact that one transformer has an iron core while another has an air core makes little difference in the mathematical treatment. The same fundamental mathematical laws apply to both. Nevertheless, it has been found that specific methods of approach are more suitable for one case than the other. Therefore, we shall separate iron-core transformers from non-iron-core transformers in our treatment. Since we begin with iron-core types, we will review the fundamental facts and equations involved in magnetics that apply to such transformers.[1]

2. Flux Density and Magnetic Intensity

The magnetization characteristics of an iron-core transformer must be analyzed in terms of the two factors that cause considerable confusion—*H*, the field intensity (or magnetic intensity) and, *B*, the flux density or magnetic induction. To gain first a qualitative concept, consider the following. When a magnetic substance like iron is placed in a magnetic field, the iron becomes magnetized because it is ferromagnetic in character. As a result of this magnetization, the number of lines of induction within it (the *flux density* symbolized by *B*) is greater than the number of lines of force (the *magnetic intensity* symbolized by *H*). The difference between *B* and *H* for any given case is determined by the characteristics of the particular sample of iron in the magnetic field. If the iron is highly ferromagnetic, the ratio between *B* and *H* will be large. If the iron is of poor magnetic quality, due to impurities, the ratio of *B* to *H* will be small. It follows, therefore, that we may describe the nature of the iron by the statement:

$$\mu = \frac{B}{H} \tag{1}$$

in which μ is a constant for a given sample and is known as its *permeability*. Thus, in any given magnetic substance there are μ times as many lines of induction as there are lines of force. The constant μ, therefore, is a symbol that represents the ease with which a material can be magnetized.

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[1] For a more detailed discussion of magnetism, see *Advanced Magnetism and Electromagnetism* edited by Alexander Schure. John F. Rider Publisher, Inc., 1959.

3. Review of Magnetic Units

Although the *mks* (meter-kilogram-second) system of measurement is superior to the *cgs* (centimeter-gram-second) system in many important ways, the *cgs* system is encountered in many modern textbooks and papers. Therefore, we shall review the details of both systems and discuss the conversion factors that permit transfer from one system to the other.

Magnetic Field Intensity (H). The intensity of a magnetic field is measured by the force it exerts on a unit pole placed in the field. *In the cgs system,* the unit of magnetic field intensity is the *oersted.* It is defined as the intensity of a magnetic field in which a unit magnetic pole is acted upon by a force of 1 dyne. *In the mks system,* the unit of magnetic field intensity is the *ampere-turn-per-meter.* This unit is based upon the field intensity inside a solenoid of N turns and L meters long, having a current of i amperes flowing through its winding.

$$H = \frac{Ni}{L} \tag{2}$$

It is evident from Equation (2) that a coil having, say, 100 turns and a length of 10 cm (0.1 meter) and a current of 1 ampere flowing in it would have a field intensity within its vacuum or air core of $H = 100 \times 1.0/0.1 =$ ampere-turns per meter.

Flux Density (B). The unit of flux density in *the cgs system* is the *gauss.* The gauss is defined as a single line of magnetic flux extending perpendicularly through an area of 1 cm². *In the mks system,* the unit of flux density is the *weber per square meter.*

Magnetic Flux (ϕ). Lines of induction are referred to as magnetic flux. A single line of magnetic flux in *the cgs system* is called a *maxwell.* Thus, a gauss is a flux density of one maxwell per square centimeter. In *the mks system,* the unit of magnetic flux is the *weber.* This unit is related to the maxwell as follows:

$$1 \text{ weber} = 10^8 \text{ maxwells} \tag{3}$$

Equation (3) permits equating the units of flux density so that we can obtain the relationship between gausses and webers per square meter. That is:

$$1 \text{ gauss} = 1 \ \frac{\text{maxwell}}{\text{cm}^2} = \frac{10^{-8} \text{ webers}}{10^{-4} \text{ meters}^2}$$

so

$$1 \text{ gauss} = 10^{-4} \text{ webers per square meter}$$

Permeability (μ). The units for permeability are obtained by solving Equation (1) for each particular system. In the cgs system, the unit for permeability is *gauss per oersted.* In the mks system, μ is measured in *webers per ampere-meter.* A useful set of conversion factors may be written:

$$1 \text{ oersted} = 1000/4\pi \text{ ampere-turns per meter} \tag{3}$$

$$1 \text{ weber} = 10^8 \text{ maxwells} \tag{4}$$

$$1 \text{ weber/meter}^2 = 10^4 \text{ gausses} \tag{5}$$

$$1 \text{ gauss-per-oersted} = 4\pi \times 10^7 \text{ webers per ampere-meter} \tag{6}$$

4. Permeability of Free Space

In the cgs system, the permeability of free space (and very closely that of air at normal atmospheric pressure) is taken as one. This means that a field intensity of one oersted produces a flux density in air of one gauss. Therefore, Equation (1) may be rewritten as:

$$B = \mu H \tag{7}$$

By convention, the permeability of free space is symbolized by μ_0 and therefore:

$$B = \mu_0 H \tag{8}$$

Thus, the permeability of free space is 1 gauss/oersted. From Equation (6) it is evident that the permeability of a vacuum in *the mks system* is $4\pi \times 10^7$ webers per ampere-meter.

5. Relative Permeability

It is often convenient to speak of relative permeability, or the ratio of the permeability of a substance to the permeability of free space. That is:

$$\mu_r = \mu/\mu_0 \tag{9}$$

in which μ is the permeability of the substance and μ_0 is the permeability of free space. Since both expressions on the right side of the equation are given in the same units, relative permeability is a pure number. For nonmagnetic materials, μ_r approaches unity; for ferromagnetic substances, it often runs up into the tens of thousands.

Neither permeability nor relative permeability is constant for any specimen of magnetic material. Permeability depends upon the magnetic history of the particular specimen and the extent to which it is already magnetized at the time the measurement is taken.

Example 1. What is the relative permeability of a piece of iron in which a flux density of 1.5 webers/m^2 is produced by a field intensity of 1000 ampere-turns-per-meter.

Solution. First find the permeability of the iron in webers-per-ampere-meter using Equation (1).

$$\begin{aligned} \mu &= B/H \\ &= 1.5/1000 \\ &= 0.0015 \text{ weber/ampere-meter} \end{aligned}$$

The permeability of a vacuum is $4\pi \times 10^7$ weber/ampere-meter. Substituting in Equation (9):

$$\begin{aligned} \mu_r &= 0.0015/(4\pi \times 10^{-7}) \\ &= 1{,}200 \end{aligned}$$

Note that transformer "iron" cores (really a high grade of silicon steel or other special alloys such as Hypersil) have relative permeabilities ranging up to 10,000. An alloy such as Permalloy (78.5% nickel and 21.5% iron) is characterized magnetically by a relative permeability of over 80,000.

6. Magnetization Curves

The inconstancy of the permeability of a ferromagnetic material is readily seen from the so-called normal magnetization curve. To obtain the coordinates for such a curve, the flux density B in the

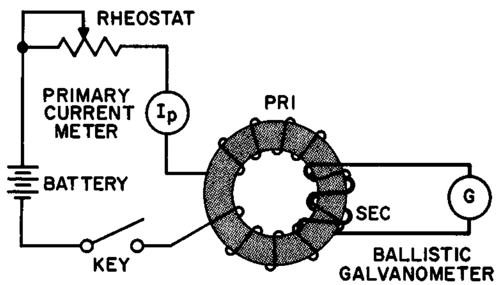


Fig. 1. Apparatus for obtaining the coordinates for the normal magnetization curve.

magnetic material is determined for various values of field intensity H. As a rule, the magnetic material to be tested is formed into a closed, doughnut-shaped toroid (often called a Rowland Ring) and wound with a primary and a secondary winding (Fig. 1).

When the key is depressed, the surge of current in the primary winding induces a secondary emf that causes a definite throw of the ballistic galvanometer. Starting with an unmagnetized specimen for each coordinate, the primary current is gradually increased (with the rheostat) and the throw noted for each value of current. The field intensity H inside the specimen may be shown to be

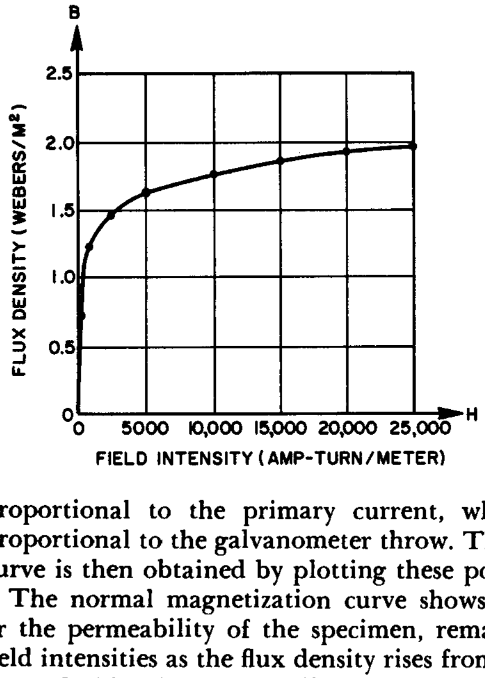


Fig. 2. The normal magnetization curve for ferromagnetic material.

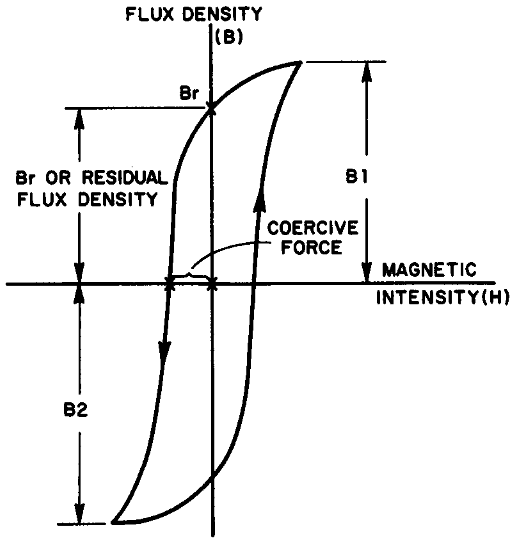
proportional to the primary current, while the flux density is proportional to the galvanometer throw. The normal magnetization curve is then obtained by plotting these points, as shown in Fig. 2.

The normal magnetization curve shows that the ratio of B/H, or the permeability of the specimen, remains constant with small field intensities as the flux density rises from 0 to about 1 weber/m^2. Beyond this, the permeability drops sharply, then gradually decreases for higher values of field intensity. As the field intensity is increased beyond 25,000 ampere-turns/meter (not shown in the graph), the specimen approaches *saturation,* a condition in which further increases in H do not yield corresponding increases in B.

Figure 3 shows the curve of flux density *vs* field intensity required to change the magnetization of a material between two values of flux density $B1$ and $B2$. A curve of this kind is known as a *hysteresis loop.* The reverse field, necessary to reduce the flux density to zero, is called the *coercive force.* The flux density remaining in the material when the positive magnetizing field goes to zero (B_r) is called the *residual flux density.*[2] The *retentivity* of the magnetic

[2] Ibid.

Fig. 3. A hysteresis loop.



substance is defined as the flux density remaining after a saturating field has acted upon it. Materials with high retentivity are called *magnetically hard* and are suitable for use as permanent magnets. If the retentivity is low, the material is *magnetically soft* and is usable as core material in chokes and transformers.

7. Core Losses

In the operation of any iron-core device, *core losses* occur in two ways. Core losses are largely responsible for the efficiency reduction of an inductive unit such as a transformer or choke and it is essential that they be kept minimal.

As alternating current passes through the windings of a transformer primary, with each reversal of current, the magnetic domains in the core material must re-orient themselves. During these current reversals, the flux density follows the hysteresis loop characteristic of the particular core. Since the domains offer opposition to re-orientation, energy that does not appear in the secondary circuit is consumed in the core substance. This energy loss is known as a *hysteresis loss.*

When the core material is soft, its retentivity is small. Similarly, the coercive force required to bring the flux density back to zero is small. Both these effects result in a diminution of the area enclosed within the hysteresis loop. If the area of the loop could be reduced to zero, the hysteresis loss would vanish, since both the residual flux density and the coercive force would disappear. Analysis has shown that the actual value of the hysteresis loss was directly proportional to the loop area.

When iron-core devices operate normally, the hysteresis loss is the same for each a-c cycle, regardless of its frequency. Thus, as the frequency increases, the hysteresis loss grows in the same proportion. Therefore, to keep hysteresis losses at a low level, the core material must be sufficiently soft so as to have a small area loop and the frequency must be relatively low. At 60 cps, the

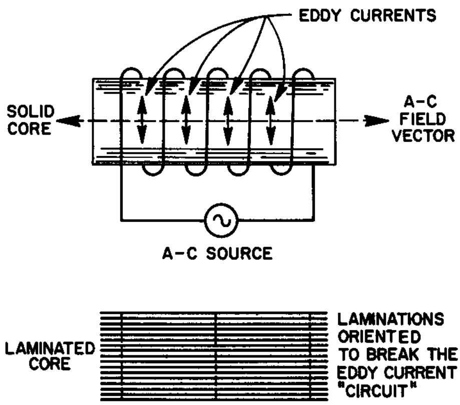


Fig. 4. A laminated core restricts the flow of eddy currents by breaking the electrical circuit at right angles to the changing magnetic field.

losses in ordinary transformer "iron" cores are tolerably low; but, at 400 cps, they may increase to the point where operation is no longer feasible. It is for this reason that special core materials are used in 400 cps equipment. The development of special core alloys such as HYPERSIL has done much to solve the problem of core losses at high frequencies. The permeability of HYPERSIL is approximately one-third higher than the usual silicon steels at comparable flux densities.

As the alternating field reverses polarity, a second type of core loss results from the flow of randomly induced currents in the core materials. *Eddy current* loss, as it is called, represents a waste of energy due to the I^2R power consumed in the core where it cannot appear as useful output in the secondary circuit. Eddy currents flow at right angles to the changing flux since maximum emf is always induced in a conductor perpendicularly to the direction of the field. The use of thinly laminated core material (each lamination is insulated electrically from the adjacent one by shellac, varnish, or an oxide scale) can reduce eddy current losses to a reasonably low figure. Because the induced currents in the core flow at right angles to the field, the laminations are always oriented parallel to the field, as shown in Fig. 4.

Both theoretical and practical considerations permit us to develop an equation whereby eddy current loss in watts can be calculated.

$$\text{Loss} = k \frac{f^2B^2t}{R} \tag{10}$$

in which f = a-c frequency in coil winding, B = a-c flux density, t = core lamination thickness, R = resistivity of the core material, and k = proportionality constant dependent upon the units used in the equation. From Equation (10), it is evident that eddy current losses in a given transformer are directly proportional to the square of the frequency, the square of the a-c flux density, and the average thickness of each lamination; and are inversely proportional to the resistivity of the core material.

8. Measurement of Core Losses

Measurement of total core loss (i.e., the sum of both hysteresis and eddy current loss) is not difficult. With the secondary circuit open, the rated primary voltage (at the rated frequency) is applied, and the power input of the transformer measured across the primary winding with a wattmeter. For this condition, the losses in the transformer consist of the primary copper loss (I^2R loss in the primary wire) and the core losses.

The primary resistance is then obtained with an ohmmeter and the primary current measured with an ammeter. The sum of the hysteresis and eddy current losses is the difference between the losses, as obtained with the wattmeter and the calculated copper loss.

To individually determine the hysteresis and eddy current losses, a power source whose voltage and frequency are both variable must be available. The procedure is as follows:

A. The sum of the core losses is obtained by the method described above, but the transformer must be loaded to its rated secondary current, so that the flux density in the core is as specified by the manufacturer.

B. The core loss at several lower frequencies but at the *same flux density* is then determined. The flux density can be maintained at a constant value, by reducing the applied voltage in the same proportion as the frequency is reduced. Since the flux density is inversely proportional to the frequency of the applied emf, reducing the voltage in step with decreasing frequency results in uniform flux density throughout the test.

C. The core loss per cycle is then plotted against frequency, with core loss handled as the dependent variable, and plotted on the Y-axis of the graph.

D. The next step involves extrapolating the curve to *zero frequency*. The core loss per cycle at zero frequency is the hysteresis loss per cycle for the particular value of B maintained during the measurement. Eddy current losses vanish at zero frequency, since there cannot be induction without a varying magnetic field. On the other hand, residual flux and coercive force still exist, so that hysteresis loss at zero frequency has a very definite meaning.

E. The hysteresis loss at the normal operating frequency is then obtained from the product of the hysteresis loss per cycle and the frequency normally used.

F. Finally, the eddy current loss is determined by subtracting the hysteresis loss at the normal operating frequency from the total core loss as obtained in Step A.

9. Current and Voltage Waveforms in Transformer Primaries

In a well-designed transformer, it may be observed that, although the applied voltage may be perfectly sinusoidal, the primary current (secondary unloaded) is far from sinusoidal in waveform. This arises from hysteresis loss effects.

Considering the primary winding of the transformer as an inductance in series with an a-c generator, we may derive the relationship given in Equation (11). e_g is the generator voltage, R is the resistance of the winding, $i\phi$ is the exciting current in the inductance, and e_i is the counter-emf developed in the coil.

$$e_g = Ri\phi + e_i \tag{11}$$

The induced voltage may be expanded, however, where N is the number of turns in the inductance, and $d\phi/dt$ is the rate of flux change in lines per second.

$$e_i = \frac{N}{10^8} \times \frac{d\phi}{dt} \tag{12}$$

If the transformer is well designed, its primary resistance will be quite small and the inductive reactance large enough to keep the induced current i_ϕ quite small. The product Ri_ϕ is normally very much less than the input generator voltage e_g. This means that the induced voltage e_i will almost be the same as the applied voltage e_g. Consequently, if e_g is sinusoidal, e_i will very closely approach the shape of a sine wave.

From Equation (12), we see that the waveform of the varying magnetic flux in the core also is nearly sinusoidal, since the two are directly proportional. However, the shape of the hysteresis loop

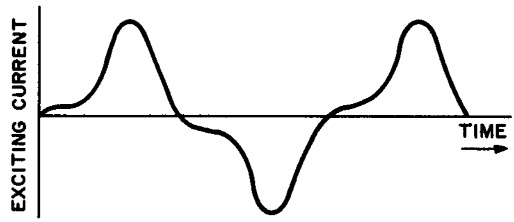


Fig. 5. Waveform of exciting current in the primary winding of a well-designed transformer.

prevents a sinusoidal flux from being produced by a sinusoidal current. Hence, the exciting current is not sinusoidal, but has a shape similar to that shown in Fig. 5. This current comprises a component that is in phase with the induced voltage and a component that lags behind the induced voltage by 90°. The in-phase current component is generally referred to as the core-loss current; the out-of-phase component is known as the magnetizing current.

10. Transformer Shielding

A transformer shield is often necessary to confine the magnetic field to a given region of space, or to prevent the effects of the field from being felt in some other area. Stray magnetic fields induce hum voltages in audio equipment, produce feedback and instability in high-frequency devices, and otherwise give rise to undesirable coupling effects.

D-c and low-frequency fields are best diverted by a shield material made of high-permeability metals. The completed shield should form an unbroken magnetic path for the flux being diverted. For this situation, the inductance of a given winding will be increased, because the shield represents a flux path of lower permeability than the air it replaces. This is especially true for high-frequency air-core coils that have little inductance in the first place.

At intermediate frequency (i-f) and radio frequency (r-f) ranges, the most effective magnetic field is made of a good electrical conductor such as copper or aluminum. The varying magnetic flux passing into the metal induces eddy currents which, in turn, give rise to magnetic flux that opposes the entering flux (Lenz's Law). The confining action in this case, therefore, arises from opposing force effects rather than from new paths created by a shield of magnetic material. An eddy-current shield should be carefully lapped and soldered (or welded) in construction, to reduce the overall electrical resistance to a minimum. An aluminum or copper shield has the opposite effect on inductance. It causes the inductance to diminish, because of the reduced permeability of the path provided by the shield.

Electrostatic shielding is generally less critical than magnetic shielding. Any metal, even one open-meshed, will, if it is carefully grounded, generally offer adequate electrostatic shielding. When

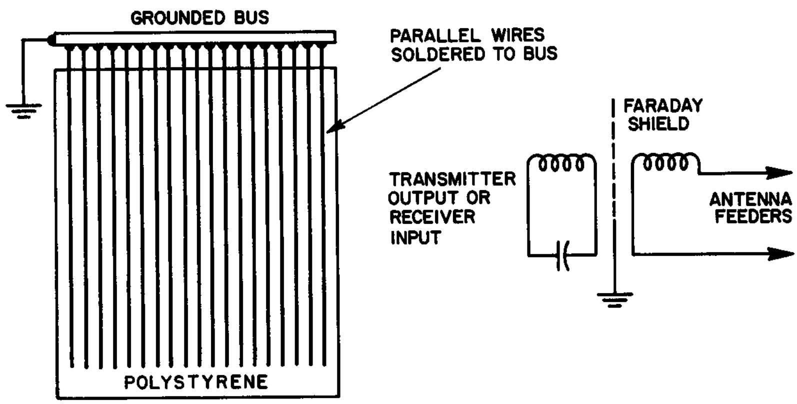


Fig. 6. A Faraday shield reduces capacitive coupling between the windings of a transformer.

individual transformer windings are to be shielded from each other, it is sufficient to wrap one layer of foil around the outside of the innermost winding and ground it at one spot. The wrapping must be cut so that there is no overlap, since overlapping would cause a shorted turn.

Frequently found between the windings of an inductively-coupled antenna system is a special type of shield known as a Faraday Shield. Since harmonics are radiated (or received) largely by capacitive coupling between the antenna coil and the resonant circuit, minimization of harmonic effects may be accomplished by arranging the coupling so that it is entirely inductive. (See Fig. 6.) A Faraday Shield must be so constructed that there are no complete electrical paths through which eddy currents can flow. A typical approach to the problem is shown in Fig. 6. Bare wires spaced their own diameter apart are cemented to a polystyrene sheet in parallel rows. Then they are connected together electrically along one end by soldering them to a length of bus wire. The bus wire is grounded to the common ground of the system on one side only.

11. Review Questions

1. Describe carefully the difference between flux density and magnetic intensity.
2. Define permeability in terms of flux density and magnetic intensity.
3. Review the meanings of the following units by defining each one and cataloging them in either the cgs or mks system:
   (a) oersted, (b) maxwell, (c) ampere-turn-per-meter, (d) gauss, (e) weber, and (e) webers-per-square-meter
4. For what medium does a field intensity of 1 oersted produce a flux density of 1 gauss? What is the permeability of this medium?
5. What is the permeability of a medium in which a field intensity of 1 oersted gives rise to a flux density of 1,500 gausses?
6. What field intensity is required to produce a flux density of $4\pi \times 10^{-1}$ webers-per-square-meter in a core material having a permeability $64\pi \times 10^{7}$ webers-per-ampere-meter?
7. What is the relative permeability of a specimen of magnetic material in which a flux density of 6 webers-per-square-meter is produced by a field intensity of 2000 ampere-turns/meter?
8. What is meant by coercive force in a hysteresis loop?
9. Name and describe the source of core losses in transformers.
10. Describe a method whereby total core loss is determined. Expand this description to cover the determination of separate hysteresis and eddy current losses.

Chapter 2

FUNDAMENTALS OF IRON-CORE TRANSFORMERS

12. Transformer Design and Construction

For low frequencies, power and communications transformers may be either of two types—core or shell. The iron core of the core-type transformer is rectangular in shape and has a clear area in the center. It may be constructed of L-shaped or I-shaped interleaved laminations, as illustrated in Fig. 7. Most transformers, especially those of the power variety, are wound so that half the primary is on one leg and half on the other leg. The same is done with the secondary winding. It is not uncommon to find other winding schemes in audio and other communications transformers. For instance, the primary and secondary are sometimes wound on the same leg of the frame, leaving the other leg free. In some cases, all of the primary winding is placed on one leg and all of the secondary on the other. [The first method minimizes leakage inductance (see Section 17) and is therefore preferred when this is a significant factor.]

A shell-type core is built up of interleaved or butt-jointed sections shaped like an E or an I. Both the primary and secondary coils are wound on the center leg of the core, with the low voltage winding going on first, i.e., nearest the core leg. Shell construction has the advantage of providing larger area flux paths that help confine the magnetic fields within the core. Most small power and audio transformers use shell-type design. Figure 8 shows the positions of the windings in a voltage step-up transformer.

Laminations are stamped from thin sheet metal. When the stamping process is completed, each lamination has a fine burr along its edges that prevents the sections from lying on one another in perfect mechanical contact. Thus, as a result of the burr making

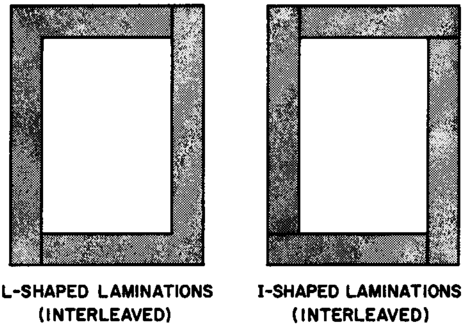


Fig. 7. Core-type transformer laminations may be interleaved L's or I's.

the effective cross-sectional area of the core somewhat smaller than its actual cross-sectional area, there is a loss of physical space. Since the space loss is a function of the method of interleaving, the transformer manufacturer is concerned with what he calls the *stacking factor* (the ratio of the effective, working cross-sectional area, to the actual or measured cross-sectional area of the core). When laminations from the same stamping die are used, a butt-jointed core without interleaving gives a greater stacking factor, and is the preferred method. A larger stacking factor means that less iron is necessary to achieve a given working cross-sectional area. For example, alternate interleaving—accomplished by alternate reversal of E's and I's—gives a stacking factor of approximately 0.87, while butt-jointing without reversal of E's and I's yields a stacking factor of better than 0.94. On the other hand, interleaving produces a core with greater mechanical strength and somewhat better magnetic-conductive characteristics. The choice of core construction depends on the exact specifications of the transformer and the particular manufacturing facilities available.

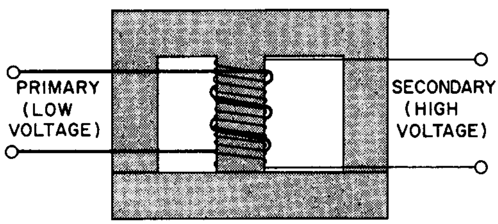


Fig. 8. Shell-type core.

13. Basic Operation Theory

It was shown that the counter-emf induced in the primary winding of a transformer (e_i) is virtually the same as the input generator voltage (e_g). The smaller the resistance of the primary winding, the more closely does e_i approach the value of e_g. In a perfect transformer, the relationship:

$$e_i = e_g \tag{13}$$

is true. From Equations (12) and (13), we may write:

$$e_g = \frac{N_p}{10^8} \times \frac{d\phi}{dt} \tag{14}$$

where N_p = primary turns

The same flux variations which affect the primary coil to which e_g is applied also act upon the secondary winding. The voltage induced in the secondary winding, as a result of the magnetizing current flowing in the primary, is then:

$$e_s = \frac{N_s}{10^8} \times \frac{d\phi}{dt} \tag{15}$$

where N_s = secondary turns

Dividing Equation (14) by Equation (15):

$$\frac{e_g}{e_s} = \frac{N_p}{N_s} \tag{16}$$

Equation (16) has been derived on the basis of two assumptions. First, that the transformer is perfect (its primary has zero resistance and the flux linkage between the two windings is 100%) and, second, that the transformer is unloaded. Since neither of these conditions is ever true in practice, Equation (16) must be considered an ideal law, subject to corrections that are implied by the actual working conditions.

Example 2. If the secondary voltage of a perfect, unloaded transformer is 12,000 volts and the primary-to-secondary turns ratio is 1:1000, what is the applied primary voltage?

Solution. Solving Equation (16) for e_g and substituting:

$$e_g = 12{,}000 \times \frac{1}{1{,}000} = 12 \text{ volts}$$

14. Efficiency and Coil Currents

Upon connecting a load to the secondary winding, the emf induced in this winding sets up a load current. Since power will now be consumed in the secondary circuit, Lenz's Law dictates that the direction of the magnetic field resulting from the secondary current must oppose the initial field from the primary. The opposing flux therefore reduces the total flux to an extent determined by the amount of secondary current flowing. Reduced flux reacts upon the primary winding by causing the counter-emf developed therein to diminish, thereby permitting a greater primary current to flow. In this way, the input to a transformer accommodates itself to match the demands made upon the output by the load. A transformer is an excellent example of an electrical device that demonstrates the wisdom of the law of conservation of energy.

A good transformer has little affect upon the power factor of the circuit to which it is connected, although small deviations from the unity power factor take on importance in special calculations. Assuming unity power factor, the efficiency of a transformer may be defined as:

$$\% \text{ efficiency} = \frac{e_s i_s}{e_g i_g} \times 100 \tag{17}$$

where e_s, i_s, e_g, and i_g denote the corresponding rms values of secondary and primary voltages, and currents.

Normally, the efficiencies of transformers run very high, with values of the order of 95% to 99% common. If the efficiency of a transformer is assumed to be 100%, then the denominator and numerator of the fraction in Equation (17) may be equated:

$$e_g i_g = e_s i_s \tag{18}$$

or

$$\frac{e_g}{e_s} = \frac{i_s}{i_g} \tag{19}$$

Substituting the turns ratio as given in Equation (16) for the voltage ratio, we have:

$$\frac{i_s}{i_g} = \frac{N_p}{N_s} \tag{20}$$

Equation 20 demonstrates that the primary and secondary currents, i_g and i_s, respectively, are inversely proportional to the number of turns on their corresponding coils.

An important point. The secondary current, i_s, is governed by the secondary voltage and the load across which this voltage is applied. Here, the secondary winding is considered to be a simple generator that obeys Ohm's Law in accordance with fundamental a-c principles. The primary current, however, is determined by the applied potential, e_g, and the impedance of the primary coil.

As shown previously, the impedance of the primary is in turn controlled by the counter-flux produced by the secondary current. Equation (20) is merely a mathematical relationship derived from prior identities that arose from physical concepts. Equation (20) does not *explain why* a given primary-to-secondary current ratio exists. As discussed below, this current ratio is a function of fluxes and counter-fluxes that are not explicitly expressed in the equation.

Example 3. What is the power output of a transformer having an efficiency of 85%, if it draws 8 amperes from a 120 volt line?

Solution. In terms of the facts given and required, Equation (17) may be restated as:

$$\%\ \text{efficiency} = \frac{\text{power output}}{e_g i_g} \times 100 \tag{21}$$

Solving the literal equation:

$$\text{power output} = \frac{(\%\ \text{eff}) \times (e_g i_g)}{100} \quad \text{so that}$$

$$= \frac{85 \times 8 \times 120}{100} = 816 \text{ volt-amperes}$$

and since unity power factor is assumed (unless otherwise stated):

power output = 816 watts

15. Coupling and Mutual Inductance

When two coils are placed near each other in a way such that a change of current in one causes a voltage to appear across the other, they are said to possess *mutual inductance.* Like self-inductance, mutual inductance is measured in henries. Two coils are said to have a mutual inductance of 1 henry, if a current changing at the rate of 1 ampere/second in one of the coils causes a voltage of 1 volt to appear across the terminals of the other coil. From this definition, it is readily apparent that the relationship between the induced voltage (e_i), the rate of change of current in one coil (di_p/dt), and the mutual inductance (M) is given by:

$$e_i = M \frac{di_g}{dt} \tag{22}$$

Self-inductance in a single coil is given by a similar equation:

$$e_i = L \frac{di}{dt}$$

in which e_i = the induced voltage, L is the self-inductance of the coil in henries, and di/dt is the rate of change of the current in amperes/second.

When two coils having self-inductances (or just *inductances*) of L_1 and L_2, respectively, are brought into a physical relationship such that the mutual inductance between them is M, it can be shown by differential calculus that the maximum mutual inductance is limited by the values of self-inductance as given in Equation (24).

$$\text{maximum } M = \sqrt{L_1L_2} \tag{24}$$

M can reach its maximum value only if the flux linkage from coil 1 to coil 2 is perfect (every line coming from coil 1 passes through

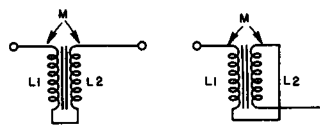


Fig. 9. Series-connected primary and secondary transformer windings.

coil 2). If the flux linkage is not perfect, M will be less than its maximum possible value. The ratio of the *actual* amount of mutual inductance to the *maximum* possible value is a useful one, because it expresses the extent to which two inductances are coupled, independently of the magnitudes of the inductances concerned. This ratio, symbolized by k is known as the *coefficient of coupling* and is defined by:

$$k = \frac{M}{\sqrt{L_1L_2}} \tag{25}$$

The coefficient of coupling is dimensionless, since both the numerator and denominator of Equation (24) are given in the same units. In a perfect transformer (*i.e.,* where there is 100% flux linkage), $k = 1$, since the mutual inductance would be maximum and hence equal to the square root of L_1L_2. This condition is sometimes referred to as *unity coupling.* Transformer windings are considered closely coupled if k is greater than 0.5. If $k = 0.01$ or less, the coils are said to be loosely coupled.

16. Experimental Determination of M and k

The primary and secondary windings of an iron-core transformer may be connected in series in one of two ways (see Fig. 9). For one method of connection, the flux generated by the primary current will be in the same direction as the flux generated by the secondary current in the common core. This situation is known as *series-aiding*. In the other connection, the two fluxes will oppose each other in the core. This is called *series-opposing*. Because a mutual inductance exists between the coils, the total inductance of the series combination will not be merely the sum of the two inductances but will include the effect of M, as follows:

$$L_{ta} = L_1 + L_2 + 2M \text{ for series-aiding} \tag{26}$$

$$L_{to} = L_1 + L_2 - 2M \text{ for series-opposing} \tag{27}$$

These relationships provide a convenient method for experimentally determining the value of M. The windings are first connected in series-aiding and the total inductance, L_{ta}, is measured. Next, they are connected in series-opposing and L_{to} is measured. (Inductance measurement may be obtained by using an inductance bridge of the correct range, or by using Ohm's Law and determining the impedance, and from this, obtaining the inductance with the help of the measured resistance of the current path, using standard a-c methods of solution.) If Equation (27) is subtracted from Equation (26), we obtain:

$$L_{ta} - L_{to} = 4M$$

or

$$M = \frac{L_{ta} - L_{to}}{4} \tag{28}$$

The mutual inductance is equal to the difference between the inductances in series-aiding and series-opposing divided by 4. Once M is determined, k is readily obtained from Equation (25).

Example 4. A transformer has a primary with an inductance of 2 henries and a secondary with an inductance of 6 henries. In series-aiding, the total inductance is 14 henries. In series-opposing, the total inductance is 2 henries. Find the mutual inductance between windings and the coefficient of coupling.

Solution. The quantities given are $L_1 = 2$ henries, $L_2 = 6$ henries, $L_{ta} = 14$ henries, and $L_{to} = 2$ henries.

Thus

$$M = \frac{L_{ta} - L_{to}}{4}$$

$$= \frac{14-2}{4} = 3 \text{ henries}$$

and

$$k = \frac{M}{\sqrt{L_1 L_2}} = \frac{3}{\sqrt{12}} = 0.866$$

17. Leakage Inductance

Even in the best transformers, not all the magnetic flux produced in one winding links with the other coil. This "leakage" flux causes an emf of self-induction and occasions what is known as *leakage inductance* in each coil. Leakage inductance behaves exactly the same way as an equivalent amount of ordinary inductance inserted in series with the winding of the transformer. Thus, it has a definite reactance and *can produce a voltage drop which increases with increasing current.*

As the load on the secondary increases, the voltage drop due to leakage inductance increases, and consequently causes the secondary terminal voltage to drop. Leakage inductance is the principal cause preventing the primary-to-secondary voltage ratio from equaling the primary-to-secondary turns ratio. In a reasonably well-designed power transformer (leakage inductance is kept low), the secondary voltage at full load should not drop more than about 8% below its no-load value. Since leakage reactance is also a function of frequency ($X_L = 2\pi fL$), the effect of leakage inductance in audio transformers is serious.

A good approximation of the total leakage inductance of a transformer may be obtained by measuring the primary inductance (with the secondary winding short-circuited) and the secondary inductance (with the primary winding short-circuited). The nearer k is to unity, the more closely will this approximation approach the true value of the total leakage inductance. If L_{mp} is the *measured* primary inductance, L_{ap} is the actual or true primary inductance. L_{ms} is the measured secondary inductance, L_{as} is the *actual* secondary inductance, and k is the coefficient of coupling, and analysis shows that the quantities are related as follows:

$$L_{mp} = 2\,(1 - k)\,L_{ap} \quad \text{with } \textit{secondary short-circuited} \qquad (29)$$

$$L_{ms} = 2\,(1 - k)\,L_{as} \quad \text{with } \textit{primary short-circuited} \qquad (30)$$

Both equations yield the expected result, if k is unity: with the linkage perfect, there is no measurable primary or secondary inductance even though the true inductances of these windings may be high. As k drops in value, the measured inductance of each winding increases. This means that the leakage inductance also increases. To keep leakage inductance low and to maintain the best possible transformer performance, k must be maintained as close to unity as possible.

The amount of leakage inductance present is *almost* independent of the core material; but, it depends upon the manner in which the coils are wound, their dimensions, their separation, etc. By using wide, very flat windings with little separation, leakage inductance is generally minimized. Interleaving the primary and secondary windings minimizes leakage inductance further.

18. Distributed Capacitance

The upper frequency response of a transformer is largely dependent upon the distributed capacitances that exist between the two ends of a given winding, between adjacent windings, and between a given winding and the core. Distributed capacitance is most often determined by direct measurement; but, it is valuable to know

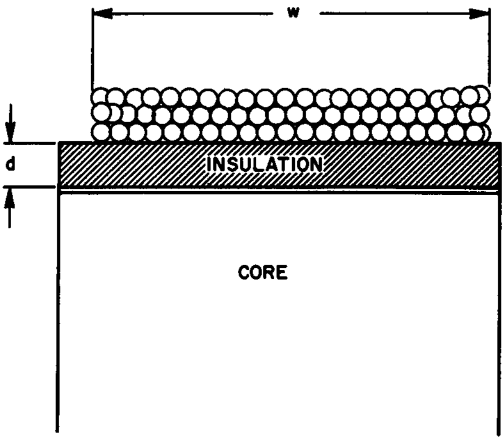


Fig. 10. Some of the factors that control the distributed capacitance between a transformer coil and its core.

the factors that determine it and how they may be controlled. Many of the equations found in transformer design literature are empirically obtained. They represent approximations that, as the number of layers in a given winding is increased, agree more and more closely with the actual measured figures. (See Fig. 10.)

Calculation of distributed capacitances is usually begun by measuring as precisely as possible the capacitance between the core and winding. (The ends of the winding are tied together for this measurement.) The equation obtained empirically from many such measurements taken is:

$$C_a = \frac{0.225\ l\ w\ K}{d} \tag{31}$$

in which C_a = capacitance between core and winding, l = mean length of a single turn in inches, w = width of the layer in inches, K = dielectric constant of the insulation, and d = thickness of total insulation between winding and core. Equation (31) demonstrates that the distributed capacitance from winding to core is directly proportional to the length of a turn (average), the width of the winding (average), the dielectric constant of the insulation, and varies inversely with the total insulation thickness.

Some circuits call for grounding one end of the winding to the chassis, which automatically grounds one end to the core. For this condition, the distributed capacitance between the core and winding decreases and

$$C_b = \frac{C_a}{3} \qquad (32)$$

where C_b is the distributed capacitance in $\mu\mu$f.

The shunt capacitance of a winding is one of the more important considerations in transformer design. It is defined as the capacitance that exists between the ends of a multi-layer coil. This capacitance, C_c, is given by:

$$C_c = \frac{0.301\ w\ K\ (N_L - 1)}{d\ N_L^2} \qquad (33)$$

where N_L = the number of layers in the coil. The distributed capacitances of a transformer change somewhat when the center is grounded (centertapped). (Exact formulas for these capacitances are empirical and may be found in any transformer designer's handbook.)

19. Transformer Sizes

The actual volume or physical size of a transformer depends on such factors as the type of core material used, the type of cooling, the permissible temperature rise for which the transformer has been designed, and the thickness of insulation material required for the potentials used. The above considerations are based on the load requirements of the transformer. However, the load requirement cannot always be considered merely the volt-ampere rating of the secondary winding. Certain usages require special handling. For example, when a centertapped transformer works into a full-wave rectifier circuit and is followed by a choke-input filter, considerable current distortion is produced in the transformer secondary winding, hence in the primary winding also. In this case, it may be shown experimentally and mathematically that for for each 100 watts of load power consumed, the average transformer dissipation rating should be about 140 watts. Thus, this type of application demands a transformer of an appreciably higher wattage rating than others in which this rectifier-filter system is used.

On the other hand, when a transformer is called upon to supply power for conventional a-c loads, its physical size is generally based upon empirical data gathered over years of engineering experimen-

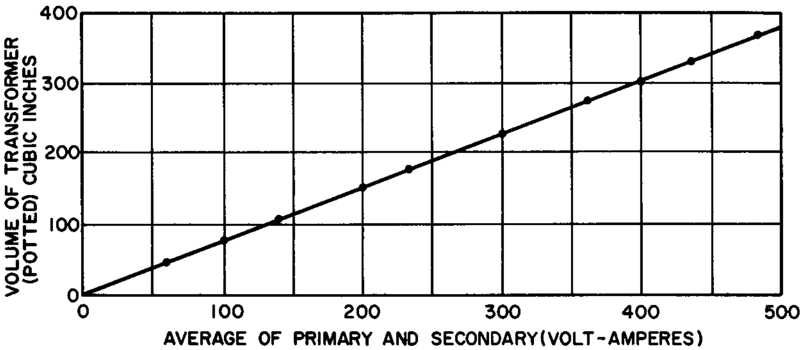


Fig. 11. Empirical size curve obtained by averaging physical sizes for several hundred transformers.

tation. In Fig. 11, an empirical curve is presented for an average transformer designed for usage in the 60 cps range, where cooling takes place only by normal convection and the allowable temperature rise is 40° C.

20. Maximum Operating Temperatures

Normal transformer engineering practice is to design for a specified temperature rise. Permissible temperature rise, in turn, is a function of the kind and quality of the insulation on the coil wires, and the insulation used to separate winding layers from each other, and from the core.

Organic materials, such as cotton, silk, paper, and varnished paper, when exposed to temperatures above certain well-defined limits tend to become dry and brittle. If the temperature continues to be high, charring results, with a further loss of mechanical strength. Overloads or vibrations under these circumstances may then produce costly breakdowns. Inorganic insulation tends to soften and melt at high temperatures. To enable the transformer engineer to intelligently choose the insulation for a given temperature rise, the American Institute of Electrical Engineers (A.I.E.E.) has promulgated a classification which establishes five grades of insulation, on the basis of maximum operating temperature. A summary of this classification is given in Table 1.

The actual temperature rise of a transformer is usually measured by determining the change in its winding resistance. A 4% resistance rise very closely equals a 10°C increase in temperature.

TABLE I

Class	*Max. Temp.*	*Materials*
0	90°C	Cotton, silk, paper, and similar organic insulators without impregnation or immersion.
A	105°C	(1) Impregnated or immersed organic insulation listed for class 0.
		(2) Molded and laminated materials with cellulose, phenolic resins, and other resins.
		(3) Films and sheets of cellulose acetate
		(4) Enamels or varnishes
B	130°C	Mica, asbestos, fiberglass, and similar inorganic materials that use organic binders.
H	180°C	Same as class B, except with silicon binders.
C	over 180°C	All mica, porcelain, glass, quartz, and similar inorganic materials.

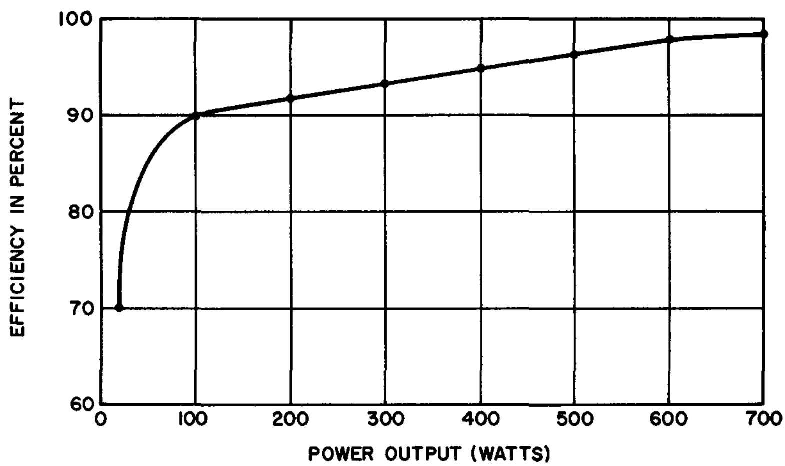


Fig. 12. Relative efficiencies of small and large power transformers, operating at 60 cps into a resistive load with a maximum temperature rise of 40°C, cooled only by convection.

21. Efficiencies of Small and Large Power Transformers

For a specified maximum temperature increase, it has been found that a small power transformer can tolerate lower efficiency operating conditions better than a large one. Since maximum efficiency is obtained when the load is so adjusted that the core losses

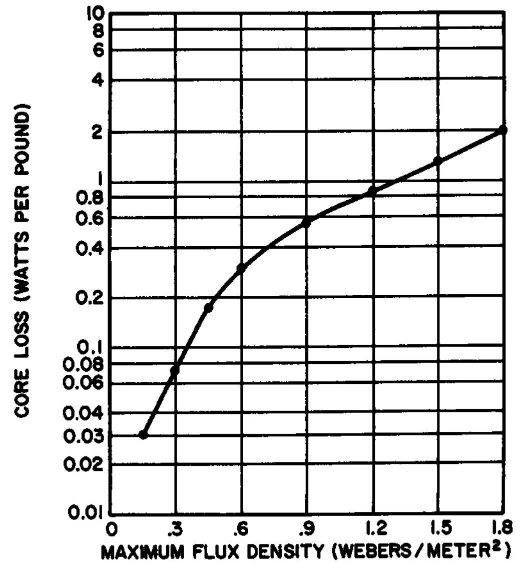


Fig. 13. Variation of core loss with increasing flux density for 60 cps core material of medium-grade silicon steel.

are equal to the copper losses, small power transformers intended for use at line frequencies below 100 cps may be designed with somewhat smaller cores than would be anticipated by considering relative power demand. In this way, less core material is used. Therefore, the efficiency of the power transformer decreases but this need not cause the temperature rise to exceed the limit set by the insulation. It is this economy in design which results in the smaller efficiencies found in low-power transformers. The comparative efficiencies of small and large transformers obtained by averaging several hundred transformers of different manufacturers are shown in the curve of Fig. 12.

22. Core Materials

Although extremely high permeability materials are now available, most commercial power transformer manufacturers still employ annealed steel laminations with a 2% to 5% silicon content.

This material possess relatively high permeability even at high flux densities. Therefore, when compared to more expensive core materials, silicon steel represents a substantial economy, since it is inexpensive and avoids excessive core losses. (Figure 13 illustrates how the core losses in a typical medium-grade silicon steel rise with increasing flux density.)

Audio transformers require core materials that have high permeability at low flux densities. This need is satisfied by alloys such as mu-metal (nickel, iron, manganese, and copper) and permalloy (nickel, iron manganese, and molybdenum).

The many grades of core materials available often make selection difficult, since there are usually several solutions to any given problem in transformer design. Generally, cost and size are the prime factors. Occasionally the choice must be based on ease of assembly, ease of mounting the completed transformer, or on the fact that only one particular core material will provide the desired electrical performance.

23. Review Questions

1. What are the differences between a shell-type and a core-type transformer?
2. What secondary voltage might be expected from a perfect, unloaded transformer if 6.3 volts is applied to a primary having 350 turns, and if the secondary contains 3500 turns?
3. Explain how the secondary load controls the amount of primary current that flows in a power transformer.
4. What is the efficiency of a 100 volt-ampere transformer if 50 volts applied to its primary causes a primary current of 2.09 amperes to flow?
5. What is meant by the equation $e_i = M\,(di/dt)$?
6. Find the coefficient of coupling if the mutual inductance between two 8-henry coils is 7.6 henries.
7. Explain fully the process and measurements that must be taken to make an experimental determination of the mutual inductance between two coupled iron-core coils.
8. What is meant by leakage inductance?
9. Explain why keeping *k* close to unity results in small leakage inductance.
10. Describe in detail how the distributed capacitance of a transformer winding is determined experimentally.
11. Discuss the various types of insulation used by modern transformer manufacturers from the point of view of transformers' maximum permissible operating temperatures.

Chapter 3

POWER TRANSFORMERS

24. Power Transformers for Electronics

Power transformers used in electronic applications are called upon to deliver alternating voltages that may be higher than or lower than or equal to the line voltage. Normally, heaters of indirectly heated tubes, or filaments of directly heated tubes, are designed to be used at voltages ranging from 1.5 to 70 volts. The vast majority of power transformers used in radio and television fall into the 6.3-volt or 12.6-volt class. These, as well as certain types of bias supplies employed in radio transmitters and industrial electronic circuits, utilize step-down transformers in which the primary-to-secondary turn ratios are proper fractions.

Plates and screens of vacuum tubes often require dc potentials higher than the peak a-c voltage provided by the line. For these applications, a plate power transformer working into a rectifier-filter system is common. Such power transformers are designed with step-up turns ratios engineered to meet the needs of the particular circuit. Plate power transformers fall into one of two classifications. They may be transformers with untapped secondary windings and used for half-wave or bridge rectification or they may contain a secondary centertap for a full-wave rectifier circuit.

Most small transformers intended for standard electronic equipment contain more than one secondary winding. For example, a power transformer used in a communications receiver might have a 500-volt centertapped secondary winding to provide 250 volts rms to each of the plates of a full-wave rectifier, a 5-volt winding rated at 2 or 3 amperes for the filament or heater of the rectifier tube, and one or more 6.3-volt windings for the heaters of the amplifiers, detector, power output tube, etc.

Some plate or combination power transformers are equipped with a tapped primary winding as well. Such transformers can be adapted more easily to locations in which the line voltage is

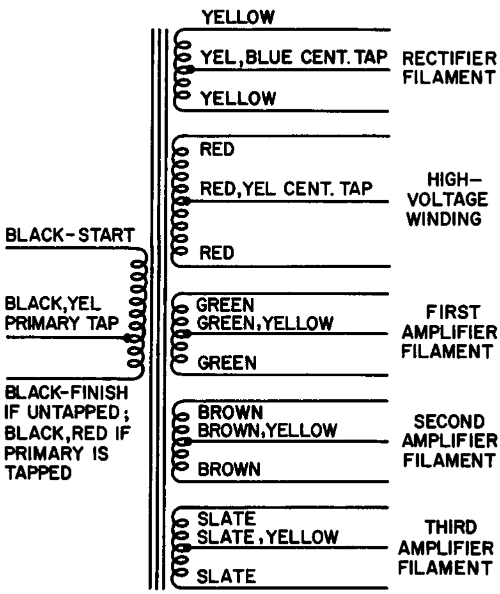


Fig. 14. Power transformer color code. This code is applicable to all types of power transformers provided with emerging, insulated wire leads, whether or not they are combination types.

slightly more or less than the nominal value given for most transformers. (The nominal line voltage used by most transformer manufacturers is 120 volts. Tube manufacturers consider 117 volts the nominal line voltage.)

In recent years, the so-called small *isolation transformer* has appeared in transformer design literature. Generally, these transformers have a single secondary winding, with the same number of turns as the primary, and are intended merely to isolate the equipment from the a-c lines, to reduce the danger of serious shock. When electronic equipment is directly connected to the lines, there is always the hazard that an ostensibly "grounded" chassis is really connected to the ungrounded leg of the a-c line. If it were, the full line voltage then would exist between the chassis and any grounded object in the vicinity.

25. Color Code for Power Transformers

Combination power and filament transformers, isolation transformers, and straight filament transformers are always color-coded, to enable convenient identification of the various windings, or they are provided with labeled binding posts or solder lugs. This is also true of plate transformers equipped with wire leads. (The applicable color code is given in Fig. 14.) The color code does not provide information relating to the actual a-c voltages available. These must be obtained from the manufacturer and generally appear on a specification sheet included in the package or on the packing carton itself. The same is also true of the current ratings of the various secondary coils.

26. Relation of Power Transformer to Rectifier-Filter System

In the normal application of power transformers in electronic equipment, filament winding ratings are always selected to meet the requirements imposed upon the transformer by the low-voltage load. For example, if a rectifier tube is specified 5 volts at 2 amperes, the filament winding should have the same rating. To be safe, it is insufficient to have a filament winding that has a higher current rating than the load. If the load does not take as much current as the winding is capable of providing at the rated voltage, then the voltage is likely to be too high. This shortens tube life. On the other hand, if the load demands more current than the rated transformer winding, the applied voltage will be too low, the transformer may begin to overheat, or both effects may occur simultaneously. Since filament windings may be obtained easily for a wide range of loads, there is no excuse for slipshod selection of the filament transformer.

The choice of output voltage for the high-voltage or plate winding, however, is not quite as simple. Since we are interested in obtaining a plate winding whose output voltage (a-c, rms) will yield a given d-c voltage at a given load current, the problem is one of first choosing a filter system and then calculating the voltage drops.

If a choke-input filter is used, the rms voltage delivered by the transformer to each rectifier diode will depend upon the desired d-c output voltage, the voltage drop in the rectifier tubes, the voltage drops in the chokes, and the current taken by the load. These quantities are related by the equation:

$$E_{rms} = 1.1 \; (E_L + \left(\frac{I\,(R_1 + R_2)}{1000}\right) + E_r) \qquad (34)$$

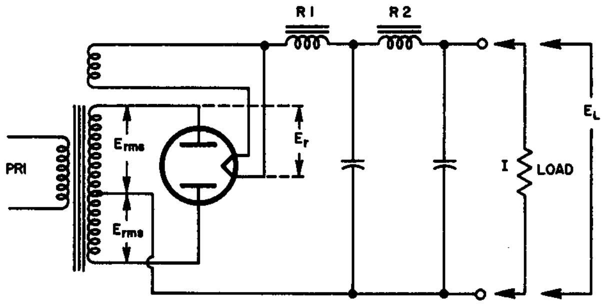


Fig. 15. Typical choke-input filter system in power supply showing voltage drops and load current.

where E_{rms} = Required rms secondary voltage of the transformer's voltage secondary to each diode. If this is a half-wave rectifier circuit, then E_{rms} is the total secondary voltage. If it is a full-wave rectifier circuit, then E_{rms} is the output voltage each side of center-tap. E_L = Voltage that is desired across the load, I = Rated full-load current (in ma), R_1 = Resistance of one choke, R_2 = Resistance of second choke, and E_r = Voltage drop across rectifier (s). (See Fig. 15.)

Example 5. Find the rms secondary voltage required to yield a d-c output voltage of 250 volts from a choke input filter having the following specifications: Choke No. 1 has a 200 ohms resistance and Choke No. 2 has a 300 ohms resistance. The load current is 50 ma and the average rectifier resistance is 1000 ohms.

Solution. Substituting in Equation (34):

$$E_{rms} = 1.1[250 + \frac{50\,(200+300)}{1000} + .050 \times 1000]$$

$$= 357.5 \text{ volts}$$

Clearly, the rms output rating of the transformer could be reduced by employing a rectifier with less internal resistance and by using chokes of lower d-c resistances.

The use of a capacitor-input filter of the type shown in Fig. 16 allows a much lower rms output rating possible for the same d-c voltage.

The equation that provides a close approximation to the rms voltage required to provide an output voltage E_L in a capacitor-input type of filter system is:

$$E_{rms} = \frac{E_L + \left[I \left(\frac{R_1 + R_2 + R_r}{1000} \right) \right]}{1.15} \qquad (35)$$

Using the values from the previous example and substituting in Equation (35), the required rms output voltage from the transformer must be:

$$E_{rms} = \frac{250 + \left[50 \left(\frac{200 + 300 + 1000}{1000} \right) \right]}{1.15}$$

$$= 283 \text{ volts or approximately } 280 \text{ volts}$$

Although there seems to be little reason for using a choke-input filter, choke-input filters are very popular. Far better voltage regu-

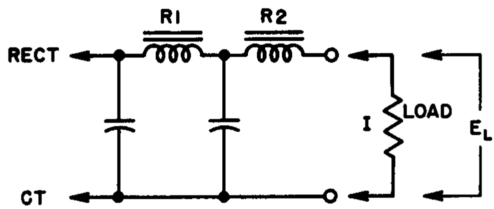


Fig. 16. Capacitor-input filter substituted for the choke-input filter of Fig. 15. Chokes have the same resistance; rectifier resistance is also the same.

lation is obtainable and higher load currents are possible with choke-input filter systems.[8]

27. Relation of Volt-Ampere Rating to Filter System

Transformers are generally designed for a predetermined maximum allowable temperature rise. When a capacitor-input filter is employed, a very high ratio of peak-to-average current may flow in the secondary winding of the transformer. This excessive ratio may result in considerably more heat in the winding than anticipated by calculating energy losses on the basis of actual power delivered to the load. This would necessitate using a transformer whose volt-ampere rating is appreciably higher than the value required for a choke-input filter. Thus, conditions very often demand the use of a choke-input filter.

In a given power supply using a choke-input filter, the choice of the first inductance in the filter system must be made on one of the following bases.

(a) If the peak rectifier current must not exceed the d-c load current, the inductance of the first choke

[8]See *Vacuum Tube Rectifiers* edited by Alexander Schure. John F. Rider Publisher, Inc., 1958.

should be related to the load resistance of the power supply:

$$L = 0.001 \times R_L \tag{36}$$

where L is in henries, and the load resistance, R_L, is in ohms. To fulfill the conditions stated, L must be equal to, or larger than, the value given in Equation (36). For example, in the illustration previously given, the power supply must provide 250 volts at 50 ma full load. Hence, at full load, the load resistance is:

$$R = \frac{E}{I} = \frac{250}{.05} = 5000 \text{ ohms}$$

To prevent the output voltage from exceeding the average voltage to the filter, the choke should have an inductance of $0.001 \times 5000 = 5$ henries, or more. This inductance value is called the *critical* value, and is especially important when the load current is small.

(b) Very often, the rectifier can handle considerably more peak current than the average d-c load current. In such cases, it is possible to use a smaller choke without endangering the tube, provided that a choke of lower inductance will still provide the desired amount of filtering.

(c) When the load current is *large,* care must be exercised to insure that the peak rectifier current does not exceed the d-c load current by too high a percentage. If the first choke has a value equal to or greater than the value given by Equation (37), the peak rectifier current will not exceed the d-c load current by more than approximately 10%.

$$L = 0.002\, R_L \tag{37}$$

Using the figures of the example previously discussed, this requires a choke of:

$$L = 0.002 \times 5000 = 10 \text{ henries}$$

This value is called the *optimum* value of inductance for a choke-input filter network.

When the optimum value is employed, conditions (a) and (c) are generally satisfied in most standard power supplies. Since the optimum value exceeds the critical value, at very low load currents—such as no drain beside that of the bleeder resistance—the output voltage will not rise above the average input voltage to the filter. When the load current rises to its maximum rated value, the inductance will still be sufficient to prevent excessive rectified peak current.

28. Filament Transformer Ratings

Most vacuum tubes in modern electronic equipment are heater types and are designed for a-c operation at relatively low voltage levels. Other than in special applications, as in high-gain voltage amplifiers employed in high-fidelity equipment, these tubes are heated directly by the a-c from a transformer. Although there has been some indication of a manufacturing trend in television receivers in the direction of series strings, it appears likely that this practice will be abandoned in favor of parallel connection of tube filaments directly across a low-voltage winding of the transformer.

A transformer of this type must be designed to carry the sum of all the currents drawn by the parallel group. In many instances, particularly in higher quality transformers, the low-voltage winding is centertapped, to permit better balance for minimizing hum.

Currents flowing in the filament or heater circuits of even medium-power tubes are, in general, quite high. Consideration must, therefore, be given to the possibility of voltage drops in the conducting wires, since under-voltage, as well as over-voltage, may be injurious to tubes. It is common practice in transmitter circuits to supply each high power filament its own filament transformer (located near the tube to avoid the long runs of connecting wires that inevitably appear when the power supply is located some distance from the other equipment). The thoriated tungsten filaments used in transmitting tubes are especially subject to deterioration, when operated at less than their rated voltages.

29. Rewinding Filament Transformer Secondaries

Filament transformers are available in many voltage and current ratings. Since they are not particularly expensive, few technicians attempt to wind or to rewind transformers. On the other hand, the appearance of transistorized equipment on the electronic scene has precipitated a demand for low-voltage transformers of all types. The service technician or engineer who finds that he requires a definite a-c voltage suitable for a testing jig, or a workbench power supply for transistors, may be interested in rewinding an old transformer to his own specifications. It is not a difficult task. It involves a small number of turns and the wire used is large enough to be handled without special precautions.

The primary volt-ampere rating of the transformer to be rewound should first be determined from the transformer label or from the manufacturer's catalog. If a secondary volt-ampere rating of 90% of the primary winding rating is anticipated, the transformer can probably be used without concern. First, marking them carefully for later identification, remove the leads from the old terminals. As the core is disassembled, note the manner in which the laminations fit together, so that they can be replaced the same way. If the primary is wound *under* the secondary, merely remove the turns from an old filament winding, counting as you go in

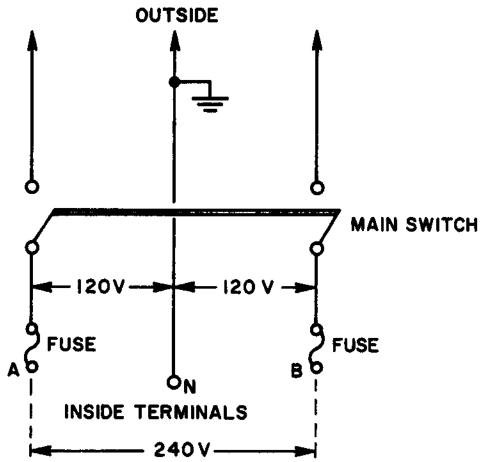


Fig. 17. Three-wire residential and factory wiring. Note that no fuse or switch is used in series with the neutral leg.

order to establish the turns-per-volt ratio. If the secondary is under the primary, the turns may still be pulled out after cutting away the side fiber insulation. The number of turns to be added in rewinding is determined by the volts-per-turn ratio as follows. Suppose that you have established the old winding turns-per-volt ratio to have been 5 turns-per-volt; assume it was a 5-volt winding . (This would be the ratio if this winding originally had 25 turns so that 25 turns to 5 volts gives a ratio of 5:1.) Should you now want a 9-volt winding, you would use 45 turns.

The wire selected for rewinding depends upon the current your load will draw. You must remain within the 90% primary volt-ampere rating limit. It is a good idea to confine your choice of wire to the smallest size that will allow 1000 circular mils/ampere. (The current ratings of various wire sizes may be obtained from any electrical or electronics handbook.) The insulation used between the core and the first winding, and between the windings themselves should be a good grade of linen paper (not friction tape), and should be wound in several layers, unless the transformer is to be used in high-voltage circuits. In that case, varnished transformer cambric should be used instead of linen paper.

30. Connection of Transformers to Power Lines

Low-power transformers, *i.e.*, 200 watts or less, used in homes and commercial establishments are simply connected to a 120-volt line, via wall or baseboard receptacles. When power requirements exceed this value, it may be necessary to take further steps to insure safe operation. The present trend in home and small factory wiring involves the use of the *three-wire* system in which 240 volts exists between the two "outside" wires, while the potential difference between each of these and the neutral, or grounded wire, is 120 volts. (See Fig. 17.) Appliances and machines are connected between the neutral and either of the two outside wires, if they are designed for 120-volt operation, or between the two outside wires (*A* and *B*) if they are heavy-duty 240-volt devices.

Observe that there is neither a fuse nor a switch in the neutral leg. If this line should be opened, full potential would exist across A and B. In addition, the voltage between the neutral and either outside line would still be present, as a result of the series connection between appliances divided between the two circuits. In practice, household appliances and small machines are divided as equally as possible across *A* and *N*, and *B* and *N* to equalize the load on both incoming "hot" wires. 120-volt rated transformer primaries are connected from either outside leg to neutral, while those of 240-volt rating are connected directly from *A* to *B*. In applications where very heavy drains are to be placed on the lines, it is often advisable to use an isolation transformer having a 2:1 step-down ratio. The primary of this transformer is connected from A to B, while its secondary steps down the voltage from 240 volts to 120 volts, for use with standard transformers. This approach results in perfect equalization of the current drain in each of the incoming wires.

31. Transformer Primary Control

Line voltages, even in industrial areas, are seldom constant throughout a 24-hour period. As the current demand of a community increases, either at the approach of evening or an unexpected storm, line voltages tend to decrease. Certain applications of transformer-operated equipment demand automatic control of primary voltage, to avoid malfunctioning as a result of fluctuating line voltage. In such applications, either voltage-regulated d-c power supplies are used or constant-voltage transformers are employed. The latter will be discussed in Chapter 6.

Where manual control of the transformer primary voltage is permissible, several methods may be used. High quality transformers having tapped primary windings are available. These enable the user to apply the input voltage to taps designed for 105 volts up to 125 volts, so that the secondary voltage is very closely equal during all normal fluctuation periods.

A second method of manual control is illustrated in Fig. 18. An inexpensive tapped secondary step-down transformer serves to boost

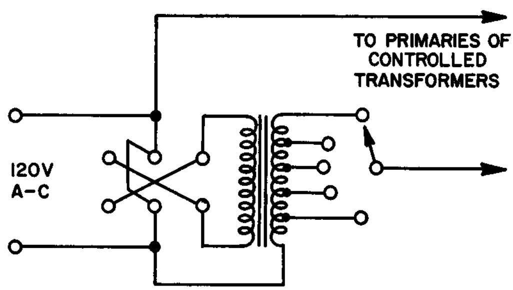


Fig. 18. Auxiliary-transformer control of input voltage to primaries of power transformers.

or buck the line voltage for adjustment purposes. The auxiliary transformer should have a secondary tapped every 2 or 3 volts over a 5- to 20-volt range. When phased in series aiding, it will add to the line voltage. When phased opposite to the feed lines, it will oppose the input voltage. A reversing switch provides for either phasing arrangement, thereby making it possible to boost or buck the line voltage at will over a 5- to 20-volt range.

The secondary should be capable of carrying the full primary current of all the controlled transformers connected to it. This method is substantially better than using a series-dropping resistor in the primary winding to reduce line voltage, since it provides for either bucking or boosting, without appreciably affecting the voltage regulation. Primary voltage control can also be accomplished by means of a variable autotransformer. The autotransformer is dealt with in Chapter 6.

32. Secondary Windings in Series

For test circuits and other special applications, low-voltage transformer secondaries may be series-connected, to increase the available voltage range. For example, a 5 volt and 6 volt pair may be connected in series-aiding to obtain 11 volts, or in series-bucking to obtain 1 volt of a-c. When using secondary windings in series, the maximum permissible current that may be drawn from the combination is limited by the winding having the lowest current rating. Proper phasing for either the aiding or bucking arrangement is determined by a-c voltage measurement, while current ratings must be obtained from the manufacturer's specifications for the particular transformer used.

33. Distribution Systems

Although a complete discussion of commercial power distribution systems is beyond the scope of this book, a discussion of power

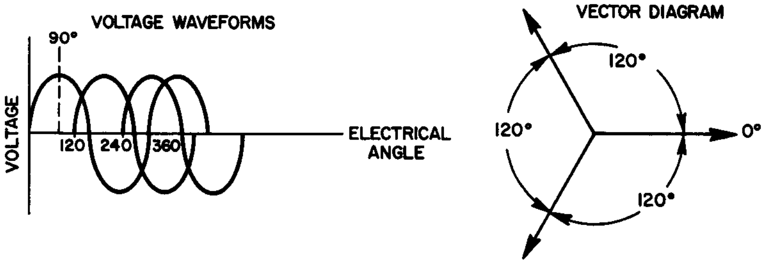


Fig. 19. Voltages in a three-phase distribution system. Each adjacent voltage is out-of-phase by 120°.

transformers would not be complete without mentioning polyphase distribution systems.

In the single-phase, three-wire system of power feed to residential and small factory electrical installations a neutral leg and two "live" legs are involved. The voltage between the two outside wires is equal to the sum of the voltages between either outside wire and the neutral. When power is derived from more than one set of armature windings in the powerhouse generator, the system is said to be *polyphase*. In general, polyphase systems are either two-phase (or quarter-phase) and three-phase. In the two-phase system, each phase is distinct from the other. The voltage between two wires of one phase always lags 90° behind the voltage between two wires of the other phase. Thus, each phase is independent of the other, and four wires are generally used for distribution. (Two-phase distribution is no longer popular.) Today, depending upon the magnitude and demands of the appliances and devices that use the electric power, single-phase and three-phase are favored.

Power in a three-phase system is derived from three sets of armature windings on the generator, and is usually distributed by three wires. The voltage across any one phase always differs from the voltage across either of the other two phases by an angle of 120°. (See Fig. 19.)

The armature sections of a three-phase generator may be connected to the feed wires that supply the loads, using either the

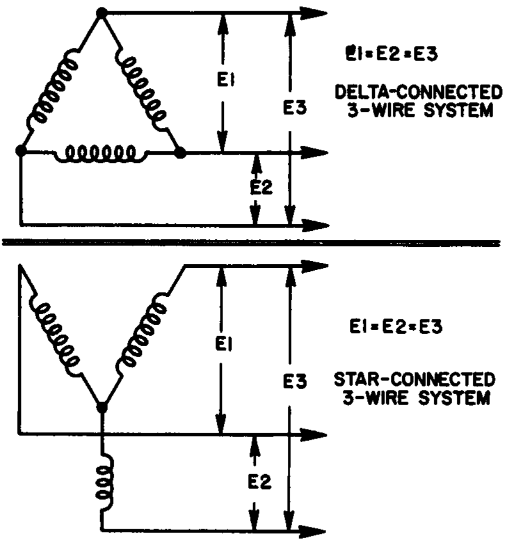


Fig. 20. Delta and star connections of three-phase generator.

delta-connection or the *star-connection*. These two methods of connections are illustrated in Fig. 20.

34. Voltage and Current Magnitudes in Three-Phase Systems

Even a cursory examination of the delta-connected generator of Fig. 20 shows that the voltage between any two adjacent wires must be the same as the voltage in each phase of the armature winding. This is also true of the voltage between the two outside wires in the delta-connection. However, the star-connection may produce a voltage between any two line wires that is 1.73 times as great as the voltage in each phase of the armature winding.

Suppose that a three-phase alternator produced 2300 volts across each of its armatures. In the delta-connection, 2300 volts of 3-phase a-c would be available between any two of the three line wires. In the star-connection, the same alternator would deliver 3980 volts across each pair of line wires. In either case, the total power delivered to the load would be the same, since the current-per-line wire would be correspondingly smaller at the higher voltage, if the current rating of the armature windings is not to be exceeded. In a star-connected generator, the armature windings carry the same current as the line wires. In the delta type, the current is divided between two armature coils. Thus, more current than that flowing in each armature coil is delivered to the line. Therefore, it follows that the current in each wire of a balanced three-phase three-wire system is equal to the current in each phase of the load if the loads are Y-connected. In the delta-connection, the current

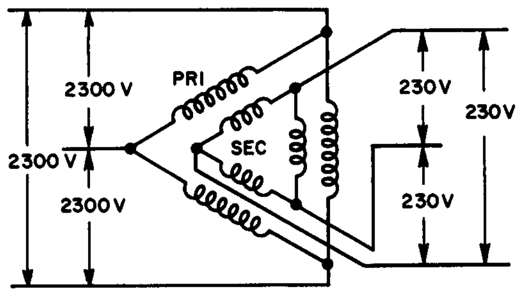


Fig. 21. Delta-delta connection of a three-phase transformer to a three-phase line.

in each wire is 1.73 times as great as the current in each phase of the load.

35. Three-Phase Transformer Connections

Transformers may be connected to a three-wire, three-phase line in several different ways. Either three single-phase transformers or one three-phase transformer may be used. Consider a distribu-

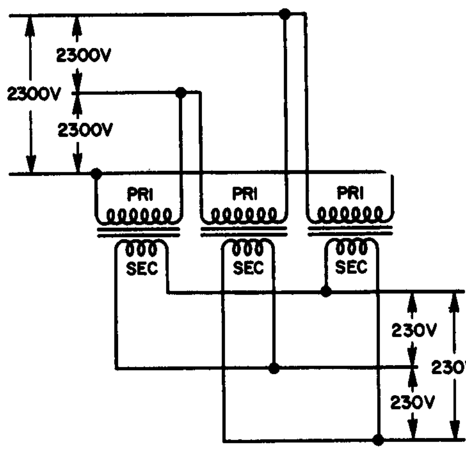


Fig. 22. Three single-phase transformers in a delta-delta connection to a three-phase line.

tion transformer system in which a 10:1 step-down ratio is to be obtained between the 2300 volt lines and the load. In this case the load would be designed for 230-volt operation. Figure 21 illustrates how a single three-phase transformer might be connected in the *delta-delta* arrangement.

In Fig. 22, three single-phase transformers are connected to a three-phase line to provide similar performance. Except for space requirements, either arrangement is capable of providing the same efficiency.

36. Review Questions

1. Explain why a power transformer intended for use with a full-wave rectifier system generally is designed with a secondary centertap. Use a diagram to help you in your explanation.
2. Why are some plate power transformers equipped with a tapped primary as well as a tapped secondary?
3. Determine the filament winding rating of a power transformer that is to supply four type 6L6G tubes with heater current. The tubes are to be parallel-connected.
4. What output voltage would be obtained from a two-section choke input filter network (choke resistances 150 ohms and 200 ohms, respectively) at a load current of 200 ma, if the power transformer rms output is 790 volts. The rectifiers are mercury-vapor types with a nominal 15-volt voltage drop.
5. Using the data given in Question 4, find the output voltage from a capacitor-input filter network with the 200-ohm choke in a single pi-filter.
6. Explain what is meant by the *critical value* of inductance in a filter network. What is meant by the *optimum value?*
7. What is the function of the centertap in the low-voltage secondary winding of a filament transformer that supplies heater voltage for high-gain amplifier tubes?
8. Explain what is meant by a "three-wire, single-phase system with one neutral." Draw a diagram to help in your explanation.
9. A transformer has two 6.3-volt secondary windings and one 2.5-volt winding. Draw a circuit diagram showing how these would be connected to obtain approximately 10 volts, indicating the phase relationships between the three windings.
10. Although the Y-Y connection of three-phase transformers has not been discussed, draw a circuit showing how this method of transformer connection would be used to supply 230 volts to a factory from a 2300 volt three-phase line.

Chapter 4

AUDIO TRANSFORMERS

37. Types of Audio Transformers

The design of audio transformers is different in many respects from the design of power transformers. The principal reason for this difference is that audio transformers must maintain a constant ratio of input to output voltage over a band of frequencies, rather than a single frequency. The problems inherent in this requirement become more severe as the width of the passband increases. (The passband is the band over which the input to output voltage remains within prescribed limits of deviation from some arbitrary reference frequency. The reference frequency will be discussed later.) Broadband design becomes more complex with the growth of the impedance levels of the transformer windings, expanded power requirements, and with the increase of d-c currents flowing in the windings. For any given type of audio transformer, the comparative design difficulty depends upon the number of octaves to be covered in operation.

INPUT TRANSFORMERS. An input transformer couples low-level input signals to the grid of the first amplifier tube or, in a multistage system, to the input element of the first transistor amplifier. Because the power level is low, to avoid induction of stray voltages in its windings, an input transformer is generally shielded. Spurious voltages would, of course, be heavily amplified under these conditions and would almost certainly appear as a distortion in the output. Generally, the principal function of an input transformer is to provide the maximum possible voltage gain while still remaining within the limits set by the system's bandwidth requirements.

INTERSTAGE TRANSFORMERS. An interstage transformer also is not required to deliver power. Its primary function is to couple the plate of one voltage amplifier to the grid of another voltage amplifier operated in class A. Since grid current never flows under class-A conditions, the transformer merely serves as a voltage step-up device. Interstage transformers are not impedance-matching devices. They must supply the maximum voltage gain possible, without reducing the frequency response of the system below the value specified by the designer.

DRIVER TRANSFORMERS. The adjective *driver* is applied only to those transformers that couple the plate (or plates) of an amplifier stage to the grids of a following class-AB2 or class-B stage. Since the grids of such output amplifier stages go positive during a portion of the cycle of the input signal, the transformer must be capable of supplying the necessary *power*. When the grids are in the negative region, there is no grid current; hence, the load resistance on the transformer secondary is quite high. As grid current begins to flow, the load resistance drops to a value determined by the magnitude of the current. This changing load reflects back to the driver tube and tends to produce serious distortion. Although this effect may be diminished by making the transformer a step-down type, when this is done, a significant secondary effect appears. A large step-down ratio places a limitation on the power that can be delivered to the driven grids. Thus, in all driver transformers, a compromise must be made between permissible distortion and required power.

OUTPUT TRANSFORMERS. Output transformers are generally employed as *impedance changers*. They change the impedance level of the output signal to that of the load. In addition, the transformer must provide d-c isolation from the load. Because output transformers always supply power, they suffer the same design problems as interstage types (often to a greater degree). Furthermore, invariably the primary winding of an output transformer carries the d-c component of the plate current of a power amplifier tube. This further complicates the design problem. We are not primarily concerned with amplifier circuits; but, to observe the differences between the various transformer types, it is helpful to study the typical circuit diagrams given in Fig. 23.

38. Equivalent Circuits

The role played by the audio transformer in an amplifier circuit may be represented in the form of equivalent circuits for low,

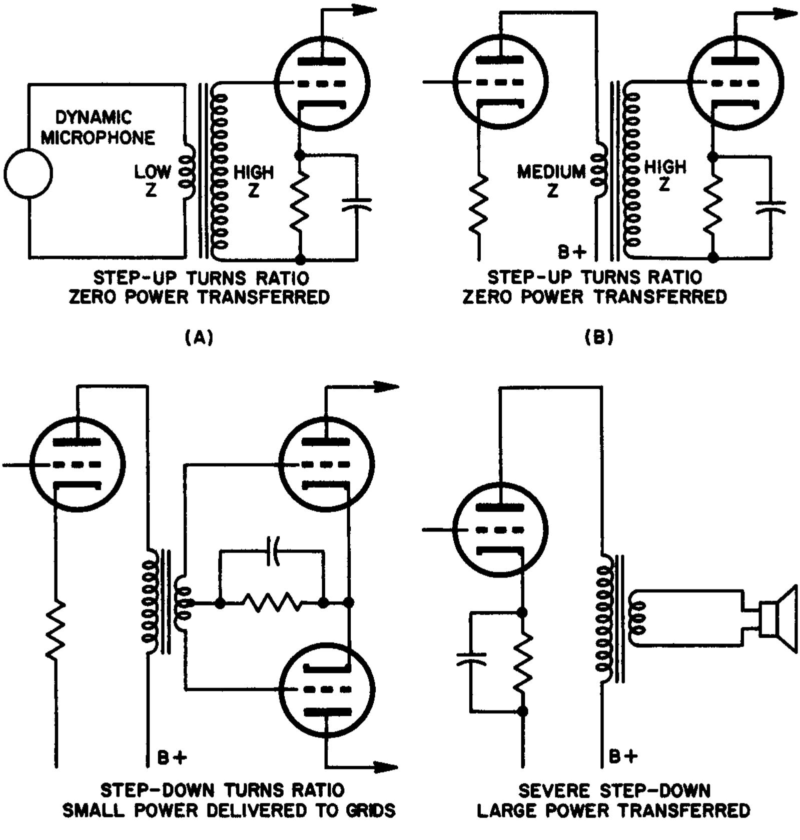


Fig. 23. A. Audio input transformer coupling a microphone to a voltage amplifier grid. B. Interstage transformer between the plate of one voltage amplifier and the grid of the next. C. Driver transformer coupling a single power amplifier to class AB2 or B grids. D. Output transformer between single-ended plate and low-impedance loudspeaker.

midband, and high frequencies. (See Fig. 24.) The limits of these frequency ranges are somewhat fluid. But it is generally accepted that the low audio frequencies range from 0 to about 400 cps and the midband frequencies range from 400 to 4000 cps. The high-frequency end stretches from 4000 cps to inaudible sound.

At low frequencies, the inductive reactance of the primary for any given transformer is relatively small, when compared to its reactance at the higher frequencies. This reactance behaves as though it were a shunt across the load and, effectively, reduces the voltage available for the load. The magnitude of this effect depends upon the transformer design. To prevent poor low-frequency response, the transformer primary must have a sufficiently large number of turns to maintain a comparatively high reactance, even when the frequency is low. This, in turn, demands that, to prevent saturation, there be a high "iron" content in the core design.

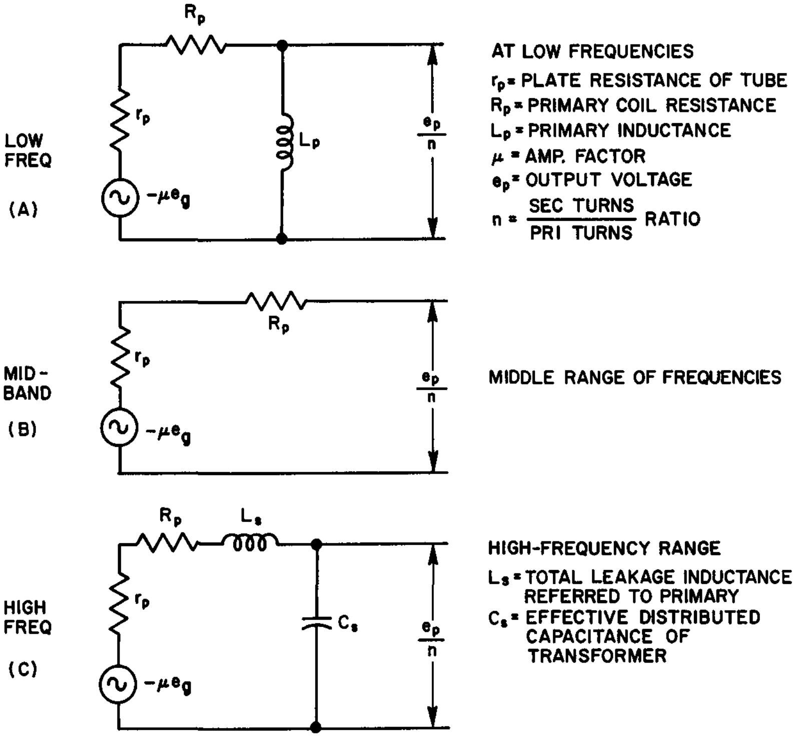


Fig. 24. A. Simplified equivalent circuit of transformer-coupled amplifier at low audio frequencies. B. Same for mid-band frequencies. C. Same for high frequencies.

Furthermore, the size wire used in winding the transformer must be large enough to keep R_p small; otherwise, the voltage drop across the resistive component of the transformer primary impedance may become large. A winding's ohmic resistance losses result in loss of signal transfer to the secondary coil. Thus, in order to meet these requirements, a high-grade transformer must be *large*. (Large is used in the comparative sense.)

In the middle frequency range, the primary reactance is high.

Hence, it does not appear as a shunting element in the equivalent circuit. Therefore, transformer response here is usually better than at the low-frequency end.

Two new factors appear at high frequencies that influence frequency response. The leakage inductance becomes significant, behaves as a series *losser* element, and reduces signal amplitude to

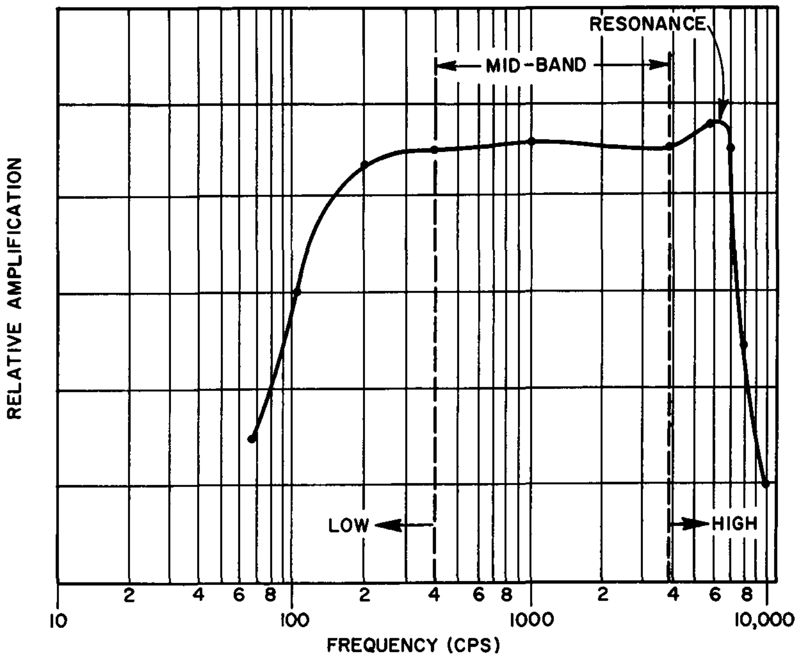


Fig. 25. Approximate response curve of medium-grade audio transformer showing dropoff at high and low frequencies, and the effect of the series-resonant circuit formed by L_s and C_s.

the output device. The effective distributed capacitance appears in shunt with the output since, at these frequencies, the reactance of the primary and secondary distributed capacitances becomes quite small—small enough to enable the capacitance to bypass some of the signal current away from the output device. At some point in the high-frequency range in most transformers, L_s and C_s reach series resonance. The result of the resonant action may, in some cases, cause a sudden rise of amplification at the resonant frequency; but, in a well-designed transformer, the magnitude of the pip just counteracts the natural drop-off that tends to occur due to C_s. In any case, high-frequency response falls off rapidly at frequencies that are substantially higher than the resonant frequency of L_s andC_s. Figure 25 illustrates an approximate curve, showing the frequency response over the entire audio spectrum of a typical medium-grade transformer. Note especially the low- and high-frequency drop-off, and the pip due to resonance.

39. Impedance Ratio of Audio Transformers

When power is transferred from one amplifier stage to another via an audio transformer, the latter becomes an impedance-matching device, as well as a coupling device. Perhaps the most familiar example of this dual function is that of a loudspeaker with a voice-coil impedance of 4 ohms, coupled to the plate circuit of an audio power amplifier through an output transformer. When manufacturers give audio ratings on power tubes, they specify the plate load impedance (or plate-to-plate load impedance for push-pull amplifiers) into which the tubes must operate to deliver the rated audio power output with the rated distortion. For example, the rated load impedance for a 6V6GT output tube with 250 volts on its plate is 5000 ohms. For this impedance, the power output is 4.5 watts and the total harmonic distortion is 8%.

For this application, a transformer having a primary impedance of 5000 ohms and a secondary impedance of 4 ohms (to match the voice coil) is required. The relationship between primary and secondary impedance, and the turns ratio of the transformer is:

$$N = \sqrt{\frac{Z_p}{Z_s}} \tag{38}$$

where N is the primary to secondary turns ratio, Z_p is the impedance of the primary, and Z_s is the impedance of the secondary winding. For the output transformer used as an example, the turns ration would then be:

$$N = \sqrt{\frac{5000}{8}} = 25.2{:}1$$

To construct the transformer, the primary is wound to give it an impedance of 5000 ohms at 1000 cps and the secondary turns then adjusted to provide the ratio given in the example.

Let us further consider the problem that arises when a class-C r-f amplifier in a transmitter is to be plate modulated by a pair of audio power tubes in push-pull.

Example 6. The modulated r-f amplifier operates with 1250 volts on its plate. When loaded by the antenna, the plate current is 250 ma. Stating the turns ratio required of the transformer, select the tubes and the modulation transformer so that 100% modulation can be obtained.

Solution. First determine the power input to the r-f amplifier.

$$P = EI = 1250 \times 0.25 = 312 \text{ watts}$$

For 100% modulation, the audio power must be 50% of the r-f power input. Thus, the audio power required is:

$$P_a = 312/2 = 156 \text{ watts}$$

Generally, the power capabilities of the tubes selected for such a job are made approximately 25% greater than the minimum requirements, to provide a suitable margin for possible losses. Increased by 25%, 156 watts become 195 watts. Therefore, we would select a pair of tubes each of which provide 100 watts of audio in the push-pull system.

Reference to the tube tables in any handbook discloses several types that can be used as modulators. Let us select a pair of 242C transmitting tubes. They are rated at 1250 volts plate and have a power output of 200 watts (for a pair in class B), into a plate-to-plate load impedance of 7600 ohms.

With the power output rating properly chosen, the turns ratio of the modulation transformer must now be calculated. For the r-f transmitting tube, the impedance that must be matched is the effective impedance of the plate circuit in normal operation. This is known as the modulating impedance:

$$Z_m = E_p/I_p \tag{39}$$

For the r-f tube of our example, the modulating impedance is:

$$Z_m = \sqrt{\frac{1250}{0.25}} \; 5000 \text{ ohms}$$

The transformer must match a primary input impedance of 7600 ohms to an output impedance of 5000 ohms. The turns ratio, from Equation (38), is:

$$N = \sqrt{\frac{7600}{5000}} = 1.232{:}1$$

40. Parallel-feed Method of Coupling

In certain applications—especially where the audio transformer is not designed to carry significant direct currents in its primary winding—it is possible to realize the advantages of transformer coupling and at the same time avoid the problems arising from d-c current saturation. As illustrated in Fig. 26, the plate current of the first amplifier in a two-stage cascade may be fed through a plate coupling resistor. A capacitor blocks the dc from the primary winding. This arrangement is satisfactory only when there is a voltage reserve in the power supply large enough to compensate

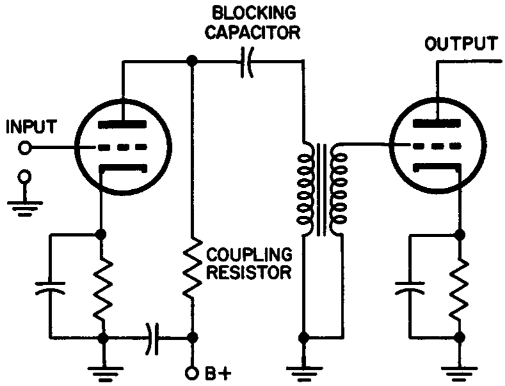


Fig. 26. Parallel plate feed of transformer-coupled amplifiers. Power supply must be able to compensate for voltage drop in coupling resistor.

for the drop that occurs in the coupling resistor. In addition, the impedance of the primary winding must be higher than ordinary transformer coupling for the same tubes requires, since the coupling resistor shunts the primary winding and reduces the effective plate circuit impedance.

41. High-fidelity Transformers

Opinions differ on the range of uniform response that constitutes the characteristic of a high-fidelity input or output transformer. A consensus of transformer manufacturers specifies a range of 30 cps to 15,000 cps. One well-known manufacturer divides the audio frequency spectrum into these categories.

Communications Range—200 cps to 3500 cps. These transformers are specifically designed for receiving and transmitting equipment, such as amateur, police, railroad, and aircraft types. The frequency response for input, output, driver, and modulation transformers is within ± 1 db over the stated voice range. A typically priced 5-watt output transformer in this group, $4.00.

Public Address Range—50 cps to 10,000 cps. These transformers are designed for typical public address applications. The frequency response is within ± 0.5

db over the entire range. A typically priced 5-watt unit in this group, $11.00.

Full-frequency Range—30 cps to 15,000 cps. Frequency response $\pm$ 1 db over the full range given. Cost for a typical 5-watt unit is $17.00.

With respect to the full-frequency range high-fidelity transformers, the manufacturer guarantees an exceptionally low percentage of distortion over the entire range both at low and high frequencies. Also included in all these units is a hum-bucking coil and core construction that provides maximum neutralization of stray magnetic fields.

Full-frequency transformers differ in design from less costly transformers in many respects. The quantity and quality of core material is generally superior, in that its hysteresis loop has a substantially smaller area. These cores do not saturate easily. The distributed capacitance of the windings is generally held to the lowest possible figure and the leakage inductance is minute.

The power output rating of an output transformer is dependent upon two factors. The maximum permissible current (which is determined by the temperature rise of the transformer windings) and the maximum permissible voltage (which is limited by the flux density in the core for the unsaturated condition).

Since the magnetization curve of any core material is nonlinear, when the flux density is high, the inductance of the winding will vary. This variation occurs within each individual cycle and gives rise to distortion. Therefore, the maximum voltage that can be applied to the transformer is a function of the maximum permissible distortion. With low-grade output transformers, distortion appearing at frequencies well within the transformer's "flat" range is not an uncommon occurrence. Therefore, if distortion is to be minimized, the flux density in the core must be kept at a low level. Thus, high-fidelity components generally contain more iron than their less expensive counterparts. This is particularly true if good response is desired at very low audio frequencies.

42. Transistor Transformers

In recent years, transistor transformers have become generally available. Usually, these units are extremely tiny, so that they may be used to advantage in miniaturized equipment such as pocket broadcast radios, portable test equipment, etc. Often, to make them impervious to moisture and changes in altitude, these transformers are encapsulated. As might be expected, the low-frequency response of such transformers is generally not as good as that found in high-quality, large transformers (see Fig. 27). But their high-frequency response is more than acceptable. The transformer whose response curve appears in Fig. 27 has a primary impedance of 1200 ohms,

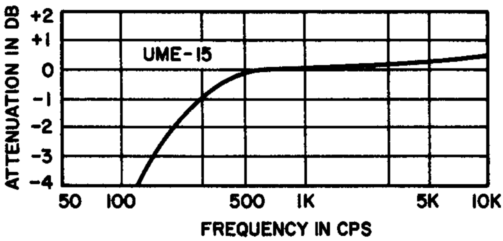


Fig. 27. Response curve of a typical ultra-miniature encapsulated transistor transformer. Courtesy, Chicago Standard Transformer Corp., transformer type UME-15.

a secondary impedance of 3.2 ohms for a miniature loudspeaker, a primary current rating of 2 ma maximum, and an output level of 100 mw.

The response curves for the remainder of this line of transformers

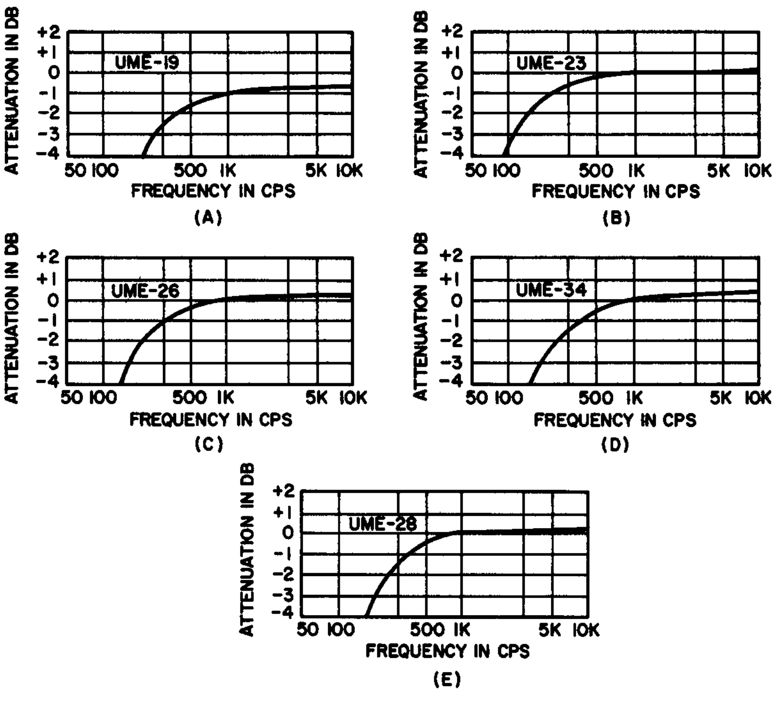


Fig. 28. Response curves of typical transistor transformers.

are shown in Fig. 28. The typical curves given in this figure are representative of transformers having the following important characteristics:

A. Output or driver transformer. Primary impedance is 10,000 ohms. Secondary impedance is 500 ohms. Power rating is 100 mw.

B. Single or push-pull output. Primary impedance is 300 ohms. Secondary impedance is 12 ohms. Power level is 500 mw.

D. Input. Primary impedance is 200,000 ohms. Secondary impedance is 1000 ohms. Power level is 25 mw.

E Single or push-pull output. Primary impedance is 7500 ohms. Secondary impedance is 12 ohms. Power level is 500 mw.

From these few examples (this particular line carries 29 distinct types), it is evident that an extremely wide range of primary and secondary impedances are available. Another complete line of transistor transformers contains 77 distinct types with impedance ranges as given below. The multitude of different types permits the selection of almost any combination of primary to secondary impedances for matching the many types of transistors in use. In the groups listed, the power levels range from 100 mw up to 350 mw.

A. Output. Primary impedance is 10,000 to 48 ohms. Secondary impedance is 500 to 3.2 ohms.

B. Driver. Primary impedance is 20,000 to 1500 ohms. Secondary impedance is 3000 to 200 ohms.

C. Input. Primary impedance is 500,000 to 3 ohms. Secondary impedance is 80,000 to 30 ohms.

Transistor transformers for power transistors may now be obtained for entertainment and experimental applications. They differ from the ones discussed above in that they have a much higher primary current rating and a higher power output rating. For example, a unit rated at 6-watt power handling ability can carry 500 ma of unbalanced d-c primary current. Its primary impedance is 48 ohms and its secondary impedance is 3.2 ohms. The frequency response is ± 2 db over the range from 70 to 20,000 cps.

43. Review Questions

1. How do audio transformers differ from power transformers in design and construction?
2. Detail the requirements for the design of input, interstage, driver, and output audio transformers.
3. Describe the behavior of audio transformers at low-, intermediate-, and high-audio frequencies, using the specific equivalent circuit in your explanation.
4. Find the primary-to-secondary turns ratio of a transformer that has a primary impedance of 7500 ohms, if it is to match a loudspeaker with a voice coil impedance of 16 ohms.
5. What is the most important advantage of parallel feed transformer coupling? What is one disadvantage?
6. Outline the frequency response requirements for a transformer designed for the communications range, the high fidelity range, and the public address range.
7. Explain why the flux density in a transformer must be kept low in order to avoid distortion.
8. Why is the frequency response in the low portion of the audio spectrum poor in transistor transformers?
9. Discuss the relative performances of the transformers whose response curves are given in Fig. 28.
10. Explain why a transformer having a hysteresis loop of small area can be expected to give better frequency response than one in which the loop has a large area.

Chapter 5

HIGH-FREQUENCY TRANSFORMERS

44. General Considerations

High-frequency transformers are most often used as coupling devices in electronic equipment where amplifier stages follow one another in cascade, or where a transfer of energy is to occur between a transmission line and a source or a load. A study of high-frequency transformers generally involves the following considerations: The transformer as a coupling impedance, the bandwidth of the coupling system, and the gain made possible by the use of the transformer. Frequently, the engineer is faced by problems that simultaneously involve all three of these factors.

Analysis of high-frequency transformers (or inductively-coupled circuits, as they are often called) involves a study of resonant and nonresonant primary and secondary windings in various combinations. Therefore, we consider examples in which neither winding is tuned, only one winding is tuned, and cases where both the primary and secondary windings are resonant. Furthermore, the behavior of high-frequency transformers is strongly dependent upon the coefficient of coupling between primary and secondary, insofar as this affects coupling impedance, gain, and bandwidth.

45. High-frequency Transformer—Neither Winding Tuned

In many respects, the analysis of an untuned high-frequency transformer is so closely analogous to that of a low-frequency iron-core transformer that the equations are the same. For example, if the coupling coefficient is unity, and if the copper losses and distributed capacitance are close to zero, the voltage ratio will be the same as the turns ratio is in a power transformer. Since all of

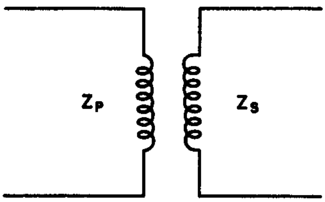


Fig. 29. Untuned high-frequency transformer.

these conditions are never satisfied in practice, the determination of the behavior of an untuned high-frequency transformer must be handled by a somewhat different method. (See Fig. 29.)

The primary and secondary windings have impedances of their own, Z_p and Z_s. The primary impedance, with zero load on the secondary (secondary open-circuited) consists of the resistance and inductance of the primary coil. When the secondary is open, the impedance of the entire secondary circuit may be considered to be infinite. For this condition, the secondary has no affect upon the primary winding and it behaves as though it were an isolated coil.

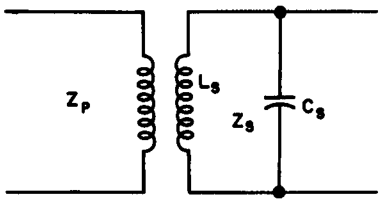
With a load connected across the secondary coil, current flows. The resulting magnetic flux opposes the current in the primary winding and causes a new impedance to appear that was not present before secondary current began to flow. The effect of this inductive opposition is equivalent to adding an impedance in series with the primary winding. This impedance is known as *coupled* or *reflected* impedance. The value of reflected impedance is given by the expression:

$$Z_r = \frac{(2\pi fM)^2}{Z_s} \qquad (40)$$

where Z_r = reflected impedance in ohms, f = frequency in cps, M = mutual inductance in henries, and Z_s is the secondary impedance in ohms. There will be little reflected impedance if the coefficient of coupling is small, since in that case M is also small. (The same is true if the impedance of the secondary is large.) As M grows, and as the load on the secondary decreases making Z_s smaller, the coupled impedance becomes significant. As a result, the voltage and current relationships in the primary are then substantially dependent upon the secondary characteristics. Also, a large transfer of energy from primary to secondary may occur under these conditions.

It is important to recognize the phase relationship between the secondary impedance and the reflected impedance into the primary.

Fig. 30. High-frequency transformer with untuned primary and tuned secondary.



Mathematical analysis shows that, although the phase angles are the same, the signs are reversed. For instance, if the impedance of the secondary circuit is inductive with a phase angle of 45°, the impedance coupled back to the primary circuit is also 45° but is *capacitive* in phase characteristics. This sign reversal has an important bearing on the behavior of high-frequency transformers. Note also that the coupled impedance will be a pure resistance, if the secondary impedance is a pure resistance. This case is particularly important when resonant circuits are analyzed since, in such circuits, the inductive and capacitive components cancel each other and leave purely resistive loads. (This will be considered later.)

To describe another special case, consider what occurs when the secondary winding of an untuned high-frequency transformer is short-circuited. For this condition, the secondary circuit resistance is negligibly small compared to the secondary inductance. Thus, the coupled resistance may also be taken as zero and the only coupled impedance is capacitive. This serves to reduce the inductance of the primary to an extent limited only by the coefficient of coupling. Thus, for a unity-coupled circuit, the primary inductance is completely neutralized.

Untuned high-frequency transformers have little practical application in coupled circuits and are rarely used in electronic equipment. They merely form the basis for a more extended study of single- or double-resonant types.

Before going on to the frequently-encountered coupling circuits, let us investigate how a coil shield or a metal chassis or panel may important bearing on the behavior of high-frequency transformers. A coil shield is a closed loop that may be considered to possess both inductance and resistance, and is inductively coupled to the coil that it shields. Since the inductance of the shield can is coupled back to the coil as a capacitance, it reduces the effective inductance of the coil. If the can has low resistance (as is true of copper and aluminum shields), the reflected resistance may be disregarded. Thus, if the coil is part of a resonant circuit, as it usually is, the resonant frequency of this circuit will increase.

46. High-frequency Transformer with Untuned Primary and Tuned Secondary

Transformers with untuned primaries and tuned secondaries are often found in cascaded r-f and i-f amplifiers. (See Fig. 30.) The coupled impedance of this circuit is given by the same equation as in the case of the transformer with neither winding tuned (Equation 40). When this equation is compared with the expres-

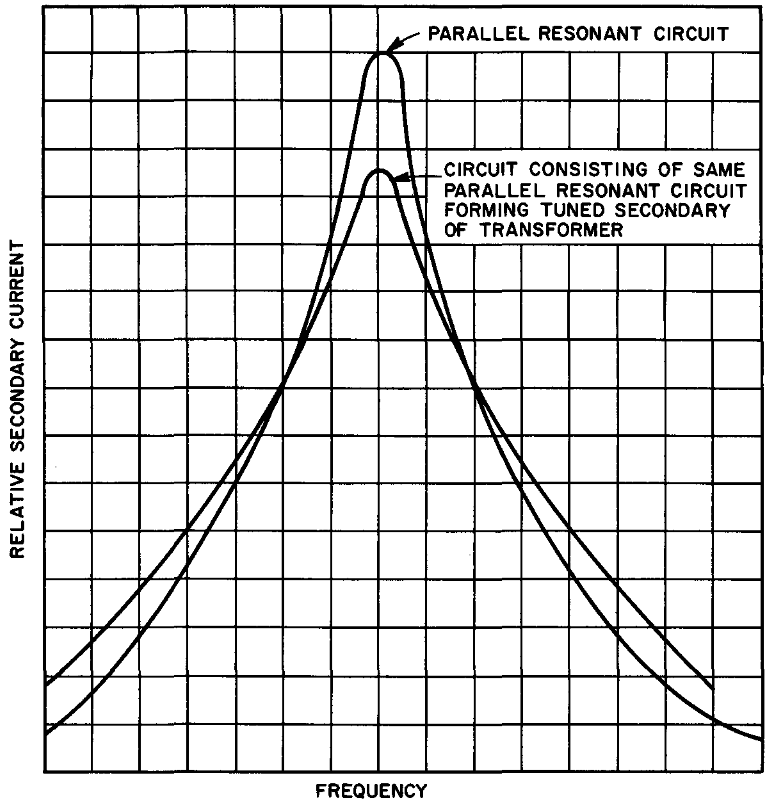


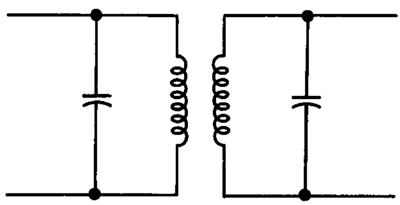
Fig. 31. Curves showing that the coupled impedance varies in a manner that is similar to the resonance behavior of a parallel-resonant circuit.

sion for a simple parallel resonant circuit, it is seen that both have the same form. Since the parallel impedance given in Equation (41) is definitely a function of frequency, it is clear that the coupled impedance is also a function of frequency.

$$Z = \frac{(2\pi f_r L)^2}{Z_s} \tag{41}$$

Since the equations have the same form, the variation of coupled impedance must be considered to vary with frequency according to the same law that governs the variation of parallel impedance

Fig. 32. A high-frequency transformer having tuned primary and secondary circuits.



of the secondary circuit alone. When this is checked experimentally, the curves obtained appear somewhat as in Fig. 31. Note that the curves have the same general form but that the *Q* of the overall transformer response (the secondary current as a function of the reflected impedance) is somewhat less than the *Q* of the secondary taken alone as a simple parallel resonant circuit. The transformer curve was obtained in a circuit in which primary source resistance was appreciably greater than impedance of the primary winding. From these curves, it is clear that a transformer with an untuned primary and tuned secondary may be used with excellent results as a selective coupling stage in electronic equipment.

47. Transformer—Both Windings Tuned

If the coefficient of coupling between a tuned primary and a tuned secondary is low (of the order of 0.007), the secondary current variation as a function of frequency follows very closely the curve given in Fig. 31 for a parallel resonant circuit, except that it may be somewhat sharper—its peak may not be as high. As the coefficient is increased, the character of the secondary current curve begins to undergo radical changes. (See Fig. 32.) In Fig. 33, we have presented four different curves, one for each of four different coupling conditions. Since actual coefficients have little meaning outside numerical problems, the qualitative terms *undercoupling, critical coupling, transitional coupling,* and *overcoupling* have been used to describe the most commonly used coefficients. The curves shown in Fig. 33 were obtained from a tuned transformer in which both windings were resonant to the same frequency. In addition, both tuned circuits had the same Q. When the two cir-

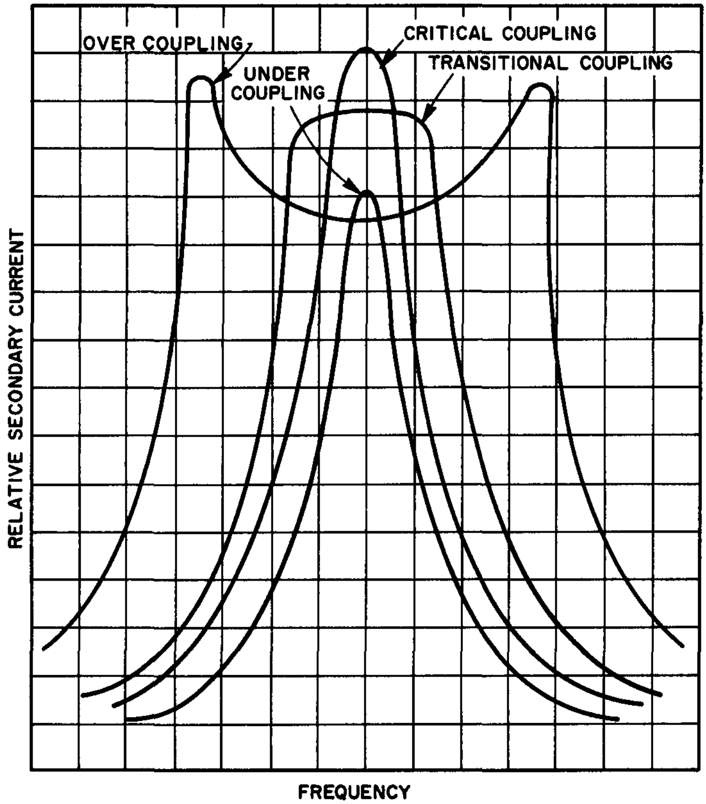


Fig. 33. Curves for various coefficients of coupling described in accepted qualitative terms.

cuits are tuned to slightly different frequencies, the curves retain the same shape, but drop somewhat in amplitude. If the Q's differ, the amplitudes of the peaks in the double-humped curves will be different.

48. Undercoupling

A tuned transformer in which both primary and secondary are resonant to the applied frequency is said to be undercoupled if, by changing the orientation of the two windings, it is possible to increase the secondary voltage. The gain of an undercoupled circuit is generally expressed as some value relative to the maximum possible gain that the circuit can provide.

Undercoupled circuits provide less gain than the circuit is capable of giving, but do have the advantage of giving the sharpest possible resonance. When i-f transformers are undercoupled, they are highly selective and are very useful in communications receivers where selectivity rather than bandwidth is important. If the coupling coefficient is much below the maximum-gain coupling, the shape of the secondary voltage response curve begins to approximate the *product* of the response curves of two circuits having Q's that are equal to the primary and secondary circuit Q's, respectively.

49. Critical Coupling

A tuned transformer is critically coupled if its windings have been oriented to obtain the maximum possible secondary voltage for a given input signal. Critically coupled circuits are encountered in the r-f and i-f stages of many high performance radio receivers.

50. Transitional Coupling

A tuned transformer is transitionally coupled if its windings have been oriented to yield the flattest possible response curve. Transitional coupling provides the widest passband without double peaks, and is often referred to as *optimum* coupling. Transitional coupling is used in radio receivers where both a large bandwidth (as compared to undercoupled or critically coupled circuits) and high-gain are required. When properly adjusted, a circuit coupled this way provides excellent all-around performance. Thus, it is desirable for low cost, medium performance superheterodyne receivers, particularly those that are not necessarily to be used for speech frequencies only.

51. Overcoupling

A tuned transformer is overcoupled if the coupling coefficient is large enough to cause the secondary response curve to show a double peak. When both the primary and secondary circuits are of very high Q, the peak amplitudes tend to be equal. For low Q circuits, the low-frequency peak is larger and the high-frequency peak is smaller in amplitude. In both cases, however, the average height of the two peaks is very nearly the same as the amplitude of the curve, when the circuits are transitionally coupled.

A wide, flat response—such as one would want in a high-fidelity superheterodyne receiver—is often obtained from a two-stage i-f amplifier, where the double-peak response of an overcoupled transformer is added to the single-peak response of a transitionally coupled circuit. The overcoupled circuit must be symmetrically tuned about the response curve of the other stage. This response can be successfully obtained only by proper selection of both *Q*'s, by careful alignment of the tuning, and by proper selection of the coupling coefficient.

52. Gain-bandwidth Factor—General

The value of an amplifier circuit is determined, in many instances, by the amount of gain that can be obtained *for a given bandwidth*. This criterion is particularly true of video amplifiers and other wideband systems where bandwidth is a primary consideration. Obviously, the gain and bandwidth of an amplifier stage are determined by the tube, the high-frequency coupling system, and the values of other significant circuit components. Many engineers refer to the *gain-bandwidth product* (*U*) of an amplifier system, considering this number to be helpful in evaluating the performance of the circuit. The gain-bandwidth product is given by the expression:

$$U = \text{voltage gain} \times \text{bandwidth (mc)}. \qquad (42)$$

Thus, if a tube is capable of producing a gain of 12 in a circuit having a bandwidth of 10 mc, its gain-bandwidth product is 120 (U = 120).

Analysis shows that *U* for a simple *resistance-capacitance* coupled amplifier may be calculated from:

$$U = \frac{g_m}{2\pi\,(C_g + C_p)} \qquad (43)$$

where g_m = transconductance of the tube, C_g = grid to ground capacitance of the following tube, and C_p = plate to ground output capacitance of the tube in question. This shows that *U* depends upon the tube's transconductance and the capacitive spacing of its elements, as long as a coupled circuit does not enter into consideration.

When transformer coupling is used between stages, the gain-bandwidth product is no longer merely a function of tube characteristics. It is strongly affected by the *Q* ratio of the two coils and by the relative values of the primary and secondary circuit capacitances. For this reason, the gain-bandwidth product as such is seldom used in inductively coupled circuits. Rather, a new element appears—the *gain-bandwidth factor.* This factor is utilized only to evaluate the performance of the coupled circuit and not the whole amplifier system.

The gain-bandwidth factor is defined as the ratio of the gain-bandwidth product of a given circuit to the gain-bandwidth product of a single-tuned circuit that has the same circuit capacitance as the sum of the primary and secondary capacitances of the coupled circuit. A single-tuned circuit is defined as a parallel combination of one inductance, one capacitance, and one resistance. Thus:

$$F = \frac{U_c}{U_s} \tag{44}$$

in which F = the gain-bandwidth factor, U_c = the gain-bandwidth product of the coupled circuit, and U_s = the gain-bandwidth product of the single-tuned circuit.

53. Gain-bandwidth Factor for Critically Coupled Circuits

The gain-bandwidth factor in critically coupled circuits is greatest when $Q_p = Q_s$, and C_p is either much larger or much smaller

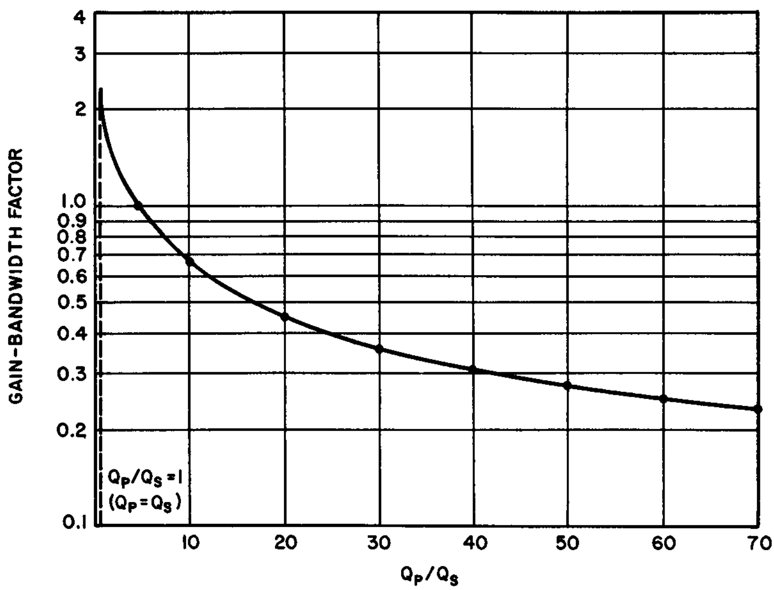


Fig. 34. Gain-bandwidth factor for critically-coupled circuit as a function of Q_p/Q_s, with $C_p = C_s$.

than C_s. If the total primary capacitance *equals* the total secondary caacitance, then even when the Q's are equal, the gain-bandwidth factor is only 1.08. This is hardly a significant improvement over the performance of a single-tuned circuit. However, if either capacitance is made 10 times larger than the other, the gain-bandwidth factor may rise to 1.88 for equal primary and secondary Q's.

The curve in Fig. 34 shows the variation of F for various Q ratios when $C_p = C_s$. To determine the gain-bandwidth factor when $C_p = C_s$, the F obtained from this curve must be multiplied by the correction given in Equation (45).

$$\text{correction multiplier} = \frac{C_p + C_s}{2\sqrt{C_pC_s}} \tag{45}$$

Example 7. Find the gain-bandwidth factor of a high-frequency, critically-coupled amplifier circuit in which the primary-to-secondary Q ratio is 4.5:1, and in which $C_p = 50\mu\mu f$ and $C_s = 10\mu\mu f$.

Solution. Referring to Fig. 34, it is seen that the uncorrected value of F is equal to 1. Applying the correction multiplier, we have:

$$F = 1 \times \frac{50 + 10}{2\sqrt{50 \times 10}} = \frac{60}{2 \times 22.4} = 1.34$$

54. Gain-bandwidth Factor for Transitionally Coupled Circuits

Transitional coupling enables the circuit designer to obtain greater gain-bandwidth factors than is obtainable with critically

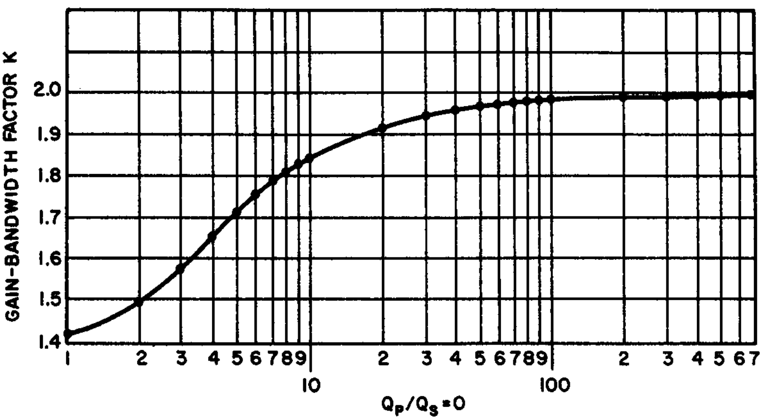


Fig. 35. Gain-bandwidth factor for transitionally-coupled circuit as a function of Q_p/Q_s, with $C_p = C_s$.

coupled circuits. The same general rules for gain-bandwidth factor determination are followed with the help of the curve in Fig. 35.

The gain-bandwidth factor is first ascertained for a given ratio of primary-to-secondary Q, assuming equal primary and secondary capacitances. If they are not, the factor obtained from the curve is then multiplied by the same correction multiplier as was used for critical coupling (Equation 45).

The curve shows that a gain-bandwidth factor range of about 1.42 to 2 is possible, with the higher values obtained when the Q ratio is large. As a comparison between the critically coupled circuit discussed in the last paragraph and a transitionally-coupled circuit having the same constants, note the solution to the following example.

Example 8. Find the gain-bandwidth factor of a high-frequency, transitionally-coupled amplifier in which the primary-to-secondary Q ratio is 4.5:1, and in which $C_p = 50\mu\mu f$ and $C_s = 10\mu\mu f$.

Solution. The gain-bandwidth factor for equal primary and secondary capacitances is first obtained from the curve of Fig. 35, and then the correction multiplier is used. For a Q ratio of 4.5:1, the gain-bandwidth factor is read from the curve as 1.69. Applying the correction:

$$F = 1.69 \times \frac{50 \times 10}{2\sqrt{500}} = 1.69 \times 1.34 = 2.27$$

Note the substantial improvement in gain-bandwidth factor for a transitionally coupled circuit as compared with an identical critically coupled circuit.

55. Gain Ratio in Critically Coupled Circuits

Gain ratio (GR) is defined as the ratio of the gain of a stage coupled in a specific manner (critically or transitionally) to the gain of a single-tuned stage having the same total circuit capacitance, without regard for bandwidth. Ordinarily, both gain and bandwidth are important considerations; under certain conditions, however, voltage gain is the prime objective. In this case, we must examine the equation for gain ratio in critically coupled circuits, to determine whether or not such circuits are capable of more gain than a single-tuned stage. The equation is:

$$GR = \frac{\sqrt{(Q_pQ_s)}}{Q} \times \frac{(C_p + C_s)}{2\sqrt{(C_pC_s)}} \qquad (46)$$

in which Q = the Q of the comparative single-tuned circuit and the remaining symbols represent the same factors as before.

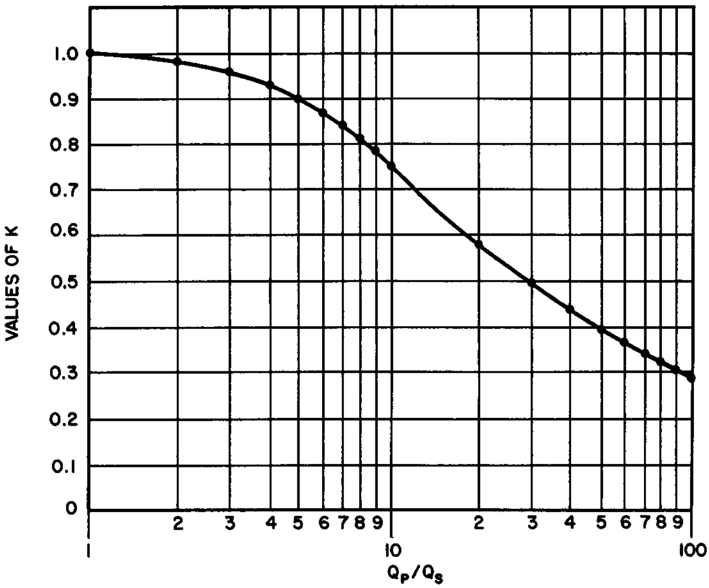


Fig. 36. The ratio k as a function of the primary to secondary Q ratio, in a transitionally-coupled high-frequency amplifier.

If $C_p = C_s$, the correction factor (extreme right-hand term) is always unity. Since $\sqrt{(Q_pQ_s)}$ is not likely to be greater than the Q of a comparable single-tuned circuit, very little benefit (with regard to gain) can be realized by using a critically coupled double-tuned circuit. To obtain a larger gain ratio in this type of stage, C_p must be made as different from C_s as the conditions permit.

To summarize, the maximum gain-bandwidth factor and the greatest gain ratio taken together will provide the best conditions for high gain and wide bandpass. These conditions can be realized by making the primary-to-secondary Q ratio as close to unity as possible, and C_p and C_s as different as possible, without controverting limitations imposed upon the system by resonance and space requirements.

56. Gain Ratio in Transitionally Coupled Circuits

The gain ratio for a transitionally coupled circuit is given in Equation 47.

$$GR = \frac{\sqrt{Q_pQ_s}}{Q} \times \frac{C_p + C_s}{2\sqrt{C_pC_s}} \times K \qquad (47)$$

where K is the ratio of the gain for transitional coupling to the gain of the same circuit for critical coupling. For all conditions except a unity Q ratio, K is less than one. Therefore, the gain ratio of transitionally coupled circuits is generally less than that of critically coupled circuits. This follows logically since, except for the multiplier K, Equation (47) is identical with Equation (46) for critical coupling. For any practical circuit in which Q_p and Q_s are known, the value for K may be obtained from the curve in Fig. 36.

Example 9. Given the circuit characteristics below, first determine the gain ratio for a critically coupled circuit and then for a transitionally-coupled circuit, as compared with a single-tuned circuit with a Q of 100.

$Q_s = 30$, $Q_p = 100$, $C_p = 50\mu\mu f$, and $C_s = 100\mu\mu f$.

Solution. For critical coupling:

$$GR = \frac{\sqrt{3000}}{100} \times \frac{150}{2\sqrt{5000}}$$

$$= \frac{54.8}{100} \times \frac{150}{2 \times 70.7} = 0.582$$

The voltage gain of a critically coupled circuit with these characteristics would be 0.582 times the gain of a single-tuned circuit.

For transitional coupling, the value of K for a transformer having a Q ratio of 100/30 or 3.33 is 0.92. Therefore, the gain ratio for a transitionally coupled circuit having is characteristics given is:

$$K = 0.582 \times 0.92 = 0.536$$

So that the voltage gain to be expected from a transitionally coupled circuit is somewhat less than that for the equivalent critically coupled circuit.

57. Review Questions

1. Explain why the bandwidth characteristics of a high-frequency transformer are often more important than its gain characteristics.
2. Define reflected (or coupled) impedance. Give the equation for reflected impedance in terms of frequency, mutual inductance, and secondary impedance.
3. Explain, in terms of coupled impedance, why the inductance of the primary of a transformer is neutralized when the secondary is short-circuited.
4. What is the principal advantage gained by using a double-tuned i-f transformer instead of a single-tuned type?
5. Distinguish carefully between undercoupling, critical coupling, transitional coupling, and overcoupling, in terms of the response curves of these systems.
6. What particular types of applications call for the use of transitional coupling? Overcoupling?

7. Define gain-bandwidth factor. How does it differ from the gain-bandwidth product?
8. Find the gain bandwidth factor of a critically coupled amplifier circuit having the following constants:

   Q ratio, P:S = 10, $C_p = 150\mu\mu f$, and $C_s = 100\mu\mu f$.
9. Determine the gain-bandwidth factor for a circuit with transitional coupling, having the same constants as those in Question 8.
10. What is meant by gain ratio? How does the gain ratio of critically coupled circuits differ from that of transitionally coupled circuits having identical constants?

Chapter 6

SPECIAL TRANSFORMERS AND APPLICATIONS

58. General

Any type of special transformer can be classified on the basis of operation frequency: power (60 cps or similar power frequencies), audio frequency, or high frequency. Discussion of special transformers in this chapter has been reserved for this time because in each case, there is something unique about the circuit, the core, or the application. Our major concern deals with the manner in which the transformer differs from standard varieties in the same group, and the special applications encountered in the field.

59. The Autotransformer

Transformer principle can be applied to a single winding system, as well as to two windings. If power is applied to the terminals of a coil, lower voltages may be obtained by transformer action from taps on the coil. If the primary voltage is made to appear between one end of the transformer and a tap on the coil, higher voltages may be obtained across the end terminals. (See Fig. 37.)

Autotransformers used in power applications are generally of the step-down variety. Referring to Fig. 38, it is evident that one section of the single winding is common to both the primary and secondary sections. Lenz's Law considerations indicate that the current flowing in this common winding must be 180° out of phase. Thus,

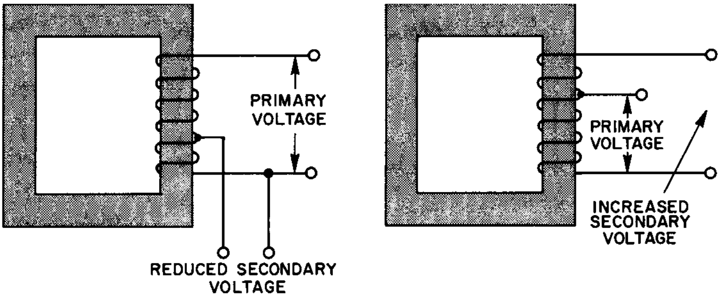


Fig. 37. Step-up and step-down autotransformer schematics.

the total current in the common section is the difference between the two. If the primary current approximates the secondary load current, the net current in the common winding may be very nearly zero. When the transformer is designed along these lines, the common section may be wound of very thin wire. This situa-

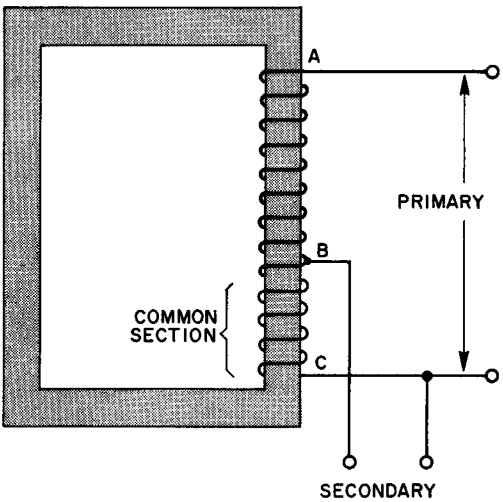


Fig. 38. The common winding may be of very thin wire under the right conditions.

tion would occur only when the primary and secondary voltages are nearly the same value, and is found in those applications where the autotransformer is used to boost, or reduce, the line voltage by relatively small amounts.

The isolation obtained between the primary and secondary of a two-winding transformer is absent in the autotransformer. This is a definite disadvantage when such isolation is desirable, as in transformers which isolate the power lines from radio chassis or other exposed metallic parts. Autotransformers cannot be used where d-c blocking is needed as, for example, in audio-coupling networks from the positive plate of one tube to the negative grid of the next tube.

Many modern television receivers utilize autotransformers in their sweep circuits. One common vertical output circuit in which

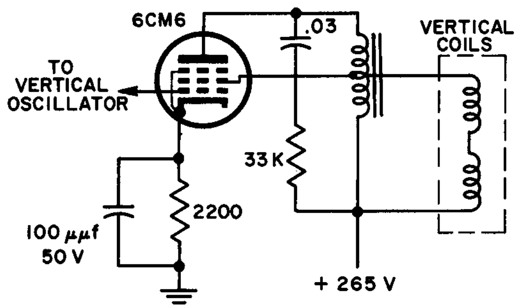


Fig. 39. Autotransformer vertical output system.

an autotransformer supplies the deflection coils with the vertical sweep current is shown in Fig. 39. This receiver also employs an autotransformer in the horizontal deflection system.

In an ideal autotransformer, the voltage ratio is the same as the turns ratio in a two-winding type. In Fig. 38, the voltage ratio would be equal to the ratio of the number of turns included between *A* and *C* to the number of turns between *B* and *C*. During operation, however, the common transformer losses reduce the output voltage to something below this figure.

60. Variable Transformers

Most variable transformers are autotransformer types in which the fixed tap is replaced by a sliding contact, to enable the user to pick off any voltage from zero up to the full line voltage. In some designs, it is also possible to obtain 10 to 15 volts more than the line voltage normally provides. This is accomplished by step-up action. (See Fig. 40.) As long as the wiper arm contact is between point *A* and point *C*, the output voltage is less than the line voltage. At point *A*, the output equals the input. If the wiper is moved to points between *A* and *B*, the output voltage is somewhat higher than the line voltage.

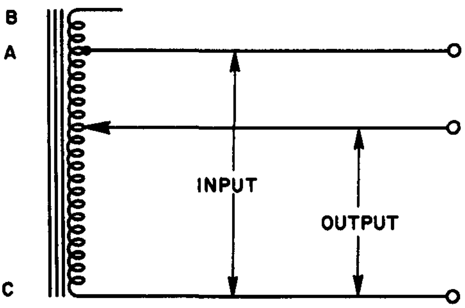


Fig. 40. Variable transformer that provides any voltage from zero to 15 volts above the line voltage.

Variable transformers are sold on the basis of both KVA ratings and current ratings. The common range for non-industrial use extends from 0.165 kva at 1.25 amperes to 19 kva at 135 amperes.

61. Instrument Transformers

The development of high-voltage alternating-current systems of transmission and distribution, necessitated the removal of various instruments from direct contact with the line circuits. Direct connection between high-tension lines and the devices on the front

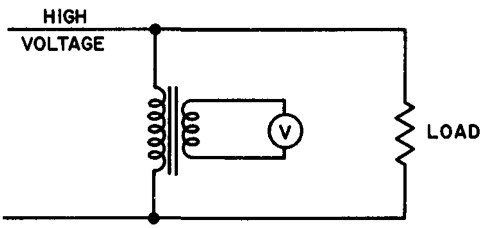


Fig. 41. A potential transformer connected to a high voltage line. The voltmeter is generally calibrated to take into account the turns ratio of the transformer.

of the switchboard must be completely avoided. By using instrument transformers, the possibility of personal injury is minimized. Transformers used for measuring high voltages at low current in this manner are known as *potential transformers.*

It is frequently necessary to meter large currents in circuits of moderate voltage. To avoid the expense and difficulty of carrying heavy leads to the switchboard, current transformers are used.

Potential transformers are generally operated under fixed conditions of applied voltage, frequency, and the number and character of the instruments in the secondary circuit. They are precision instruments and are more permanent than the instruments they operate. Figure 41 illustrates the connection method of a high voltage line, through a potential transformer, to the voltmeter.

Current transformers are sometimes called series transformers because of the way they are connected in the line (see Fig. 42). When a current transformer is in use, it is important that the secondary winding be kept closed (i.e., the instrument must not be disconnected). If the secondary is open, there will be no demagnetizing effect due to the secondary flux and, as the primary

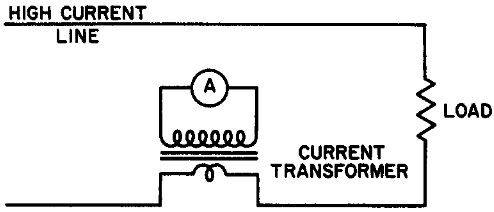


Fig. 42. The current transformer positioned to determine current flow in a heavy-duty circuit.

current is fixed by the load on the line, the total flux may rise to a high value. This tends to increase the iron losses to such an extent that the insulation may be injured by the heat. At some subsequent time, this may result in insulation puncture even for a moderate voltage. In addition to this, the secondary voltage may be very large because of the high primary current. This may lead to damage of the secondary insulation. Also there is the possibility of severe shock from the secondary terminals.

62. Saturable Reactors

Although the words *saturable reactor* imply a single coil, most saturable reactors are really two-winding devices. The inductance of a choke or reactor is a function of the degree of core saturation, as well as the number of turns and type of core material. In a saturable reactor, a d-c winding that is distinct from the main reactor, governs the extent to which the core is saturated. A typical core and winding configuration is shown in Fig. 43. Although the design of an efficient saturable reactor requires a thorough understanding of the principles of iron-core chokes carrying direct current, a general understanding of the action can be realized on a qualitative basis.

The presence of a d-c winding suggests that ac flowing in the coils on the same core might induce a current in the d-c winding by normal transformer action. This would represent an appreciable loss of power under certain circumstances. To avoid induction from the ac into the d-c coil, the core and winding directions are arranged so that the a-c fields in the portion of the core on which the d-c winding appears tend to cancel each other. The arrows in Fig. 43 indicate the instantaneous directions of the fields in the center leg of the shell. The total field due to the ac has the same direction in all parts of the large rectangle of the shell, and the

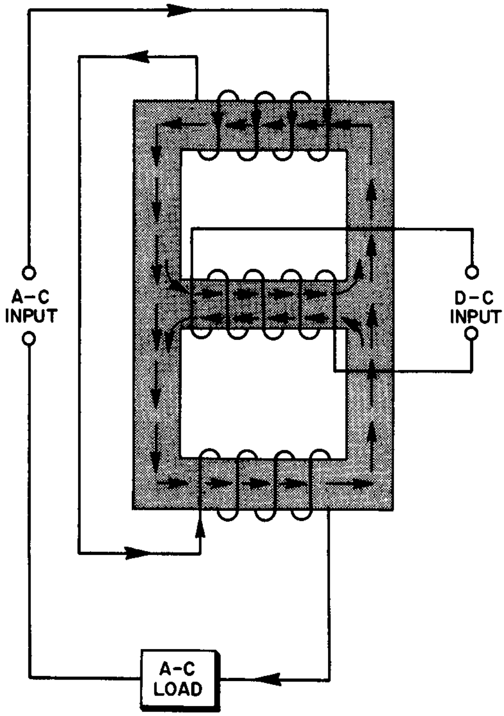


Fig. 43. A saturable reactor configuration.

fields due to the individual a-c coils have opposite directions in the core section that carries the d-c winding.

Operationally, the magnitude of the direct current is controlled by a rheostat or potentiometer. With no current in the d-c winding, the inductance of the a-c section is maximum. Hence, the inductive reactance is also at its highest value. As the d-c increases, the core saturates more easily, and the inductance and reactance decrease as well. Thus, a saturable reactor is suitable wherever there is a need for an adjustable reactor. For example, the power provided to operate a load may be varied, by changing the inductance of a saturable reactor in series with an a-c power source and the load. Saturable reactors distort the a-c waveform applied to the load. Their use is limited, therefore, to applications in which waveform is of no importance.

63. Self-saturating Saturable Reactors

When small changes in the d-c control current can produce larger changes in the a-c load current, amplification results. This action is encountered in magnetic amplifiers. Appreciably more gain may be realized, however, by employing some form of regenera-

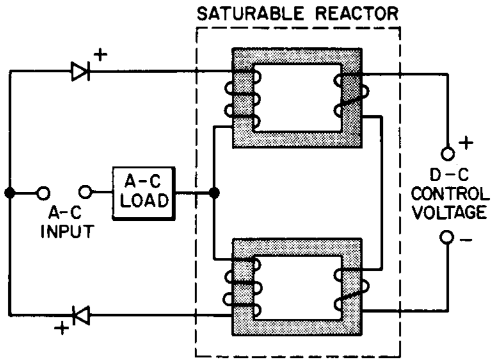


Fig. 44. Regenerative magnetic amplifier in which the load is an a-c operated device or devices.

tive feedback. Figure 44 shows one form of a regenerative saturable reactor, generally termed a self-saturating type.

An additional control winding feeds back a portion of the alternating-current output, by rectification of the load current. With regeneration, the rectified load current flows through the feedback wiring, so as to aid the d-c control current, thereby providing more gain than is possible with the simpler form of saturable reactor. An increase in gain obtained this way, unfortunately, is also associated with a rise in response time and a decrease in linearity. Often, compared to the advantages that result from the augmented gain, these factors are unimportant. In applications where response time and linearity are important considerations, degenerative rather than regenerative feedback must be used. As in vacuum tube and transistor amplifiers, by introducing degeneration, linearity is improved; but the gain is reduced.

64. Voltage-regulating Transformers

Power transformers used in radio receivers, television sets, and other similar constant-load devices are generally not equipped for regulating voltage. In other types of equipment, such as class-B modulators, the load current may swing over an extremely wide range, and tendto produce serious fluctuations in the output voltage of the power supply. Electronic voltage regulators for the heavy d-c load found in even a medium-power class-B system are difficult to build and extremely costly. In such cases, voltage-regulating transformers may be employed. Figure 45 illustrates the fundamental circuit involved in the applications of voltage-regulating transformers to a d-c power supply intended for varying loads.

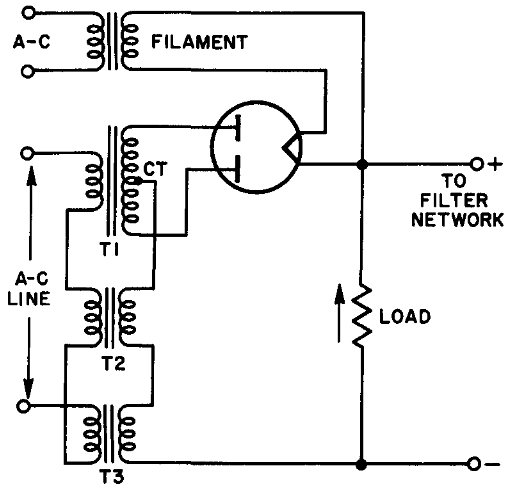


Fig. 45. Voltage-regulating transformers in a d-c power supply. **T1** is the main power transformer. **T2** and **T3** are matched voltage-regulating transformers.

The secondary coils of both regulating transformers (*T2* and *T3*) are in series with the load current from the output of the power supply. Should the load current suddenly increase, the flux density in these transformers rises, bringing them closer to saturation. This reduces the primary inductance, hence, the inductive reactance, so that the voltage applied to the primary of the main power transformer increases. Thus, the secondary voltage, which tends to drop because of the increasing load current, is brought back to its initial value by the increased primary voltage.

The output of the regulating transformers must be eliminated, hence, the primaries of *T2* and *T3* are connected in series-opposing, so that these voltages cancel. (Series opposing is indicated in Fig. 45.) If cancellation is not perfect, an a-c ripple may be introduced in the power supply output, since the rectifier tube is not associated with the regulating circuit. A successful system of this kind depends upon perfectly matching both transformers.

65. Balancing Transformers (Baluns)

The need for balancing transformers arises most often in coupling radio-frequency transmission lines to either transmitting or

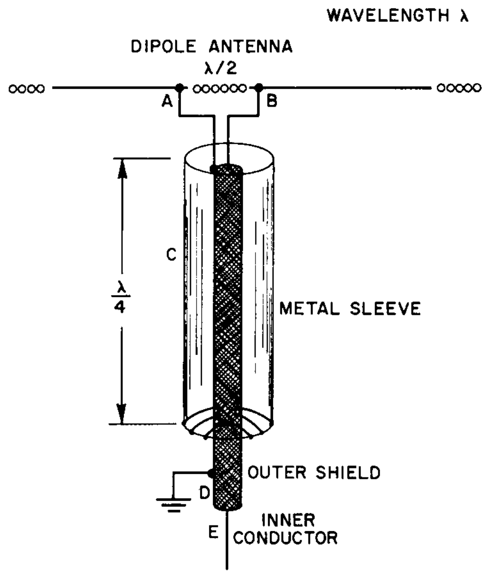


Fig. 46. A simple balancing device or **balun.**

receiving antennas. If a symmetrical centerfed antenna is connected to a balanced transmission line, such as in a parallel-conductor open wire system, the inherent balance of the entire arrangement is not upset. Remember that the absence of radiation from a transmission line—an important requisite of an efficient antenna network—occurs because of cellation of r-f currents that flow down one wire and up the other in the transmission line.

For example, if a coaxial cable is used to feed the center of a half-wave transmitting antenna, unless balancing precautions are observed, radiation from the line will occur. This is understandable when we consider that the outer sheath of the coaxial cable is a shield that does not allow the radiation from the inner conductor to reach the skin-effect rf flowing along the outside of the sleeve. Without this cancellation, radiation takes place from the sleeve, with consequent loss of power transferred to the antenna.

Balancing the system involves inserting a balancing device commonly known as a *balun* (*bal*ance to *un*balance). The simplest form of balun is shown in Fig. 46. This consists of a metallic sleeve fitted over the transmission line, and short-circuited to the coaxial shield at the bottom end. The sleeve length is made quarterwave for the frequency to be transmitted.

The input impedance of a short-circuited quarterwave section approaches infinity. This characteristic is an important factor in the performance of the balun. The explanation of balun operation is somewhat easier to understand if we consider the voltage applied from the balanced antenna side, as would be the case if the antenna were for receiving purposes. This explanation is equally valid for transmitting antennas. Consider a voltage applied to *A* and *B* at the top of the line. The two halves of the antenna atempt to establish equal peak voltages, with respect to ground potential, at these points, since both halves are symmetrical with respect to ground. This appears impossible, since the extension

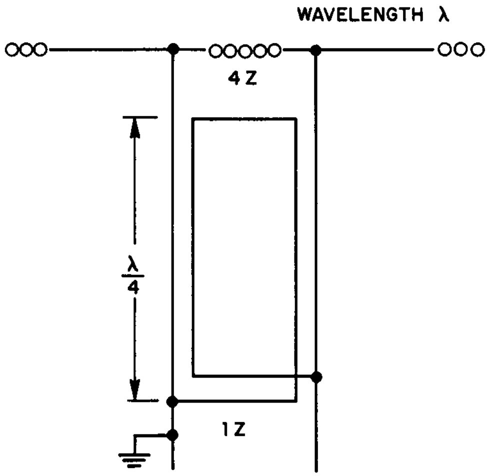


Fig. 47. A series-parallel balun arrangement used to match a balanced line to an unbalanced line. Such a balun also offers an impedance transformation of 4:1.

of A (*i.e., D*) is grounded. The sleeve *C* may be considered an extension of the outer shield of the coaxial cable *D*; but the impedance across *A* and *C* is extremely high (approaching infinity) because this is a short-circuited quarterwave section. Thus, sleeve *C* acting as an extension of *D* remains at ground potential, *leaving A free to take on any potential the antenna delivers to it.* It is clear, then, that the peak potentials across *A* and *B* can deviate from ground potential by equal amounts, despite the electrical continuity between *A* and *D*.

A second form of balun may take on a coil as well as a linear form. Transmission lines (each one quarter wavelength as shown in Fig. 47) are connected in parallel at one end and in series at the other. The transmission lines are balanced with respect to ground at the series end (top in Fig. 47). One side of the line at the other end may be grounded because the two ends are decoupled by the short-circuited quarterwave (or any odd multiple of a quarterwave) section. At the series-connected end, the line will match an impedance equal to twice the characteristic impedance of the lines used. At the parallel-connected end, there will be an impedance match to one-half the characteristic impedance of the lines. Hence, the impedance transformation from parallel end to series end is 1:4. This is a very practical way to match a 300-ohm balanced line to a 75-ohm unbalanced coaxial line.

66. Review Questions

1. How does an autotransformer differ from a standard form of transformer? What are its advantages and disadvantages?
2. Draw a diagram showing how an autotransformer would be connected between an a-c line and a load to obtain step-up action.
3. Explain the salient requirements for a current transformer and for a voltage or potential transformer.
4. With the aid of a diagram, explain how a saturable reactor would be used as a magnetic amplifier.
5. Draw a circuit diagram showing how a saturable reactor would be connected with one winding providing degenerative feedback.
6. How does a voltage-regulating transformer operate? Explain in detail.
7. Describe a sleeve type of balun and explain its operation.
8. Repeat this procedure for a linear series-parallel balun.

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