Audio Transformers

by
Bill Whitlock

Jensen Transformers, Inc.
9304 Deering Avenue
Chatsworth, CA 91311
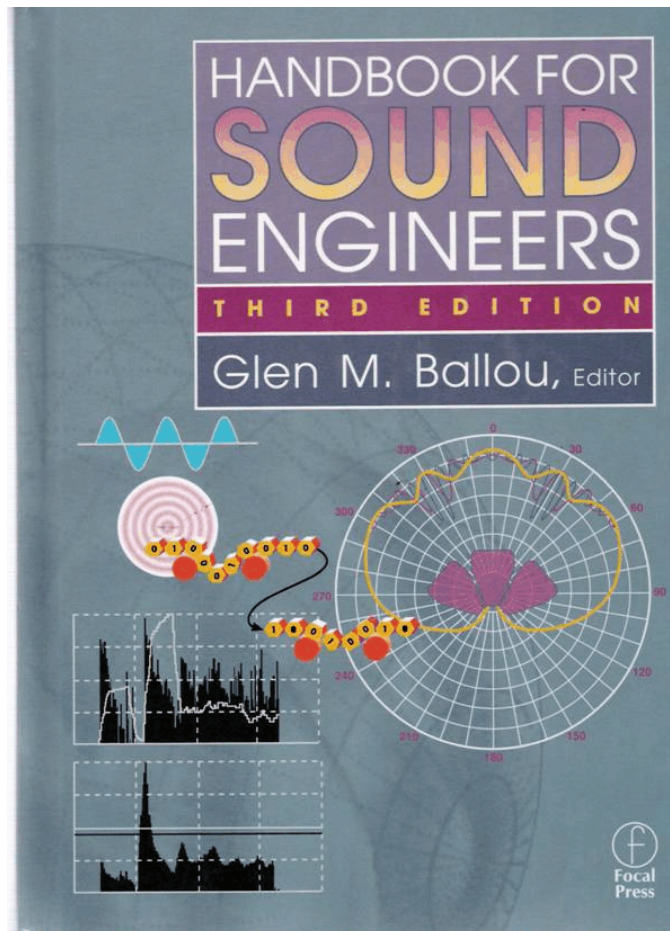
1 Audio Transformer Basics

Since the birth of audio electronics, the audio transformer has played an important role. When compared to modern miniaturized electronics, a transformer seems large, heavy, and expensive but it continues to be the most effective solution in many audio applications. The usefulness of a transformer lies in the fact that electrical energy can be transferred from one circuit to another without direct connection, and in the process the energy can be readily changed from one voltage level to another. Although a transformer is not a complex device, considerable explanation is required to properly understand how it operates. This chapter is intended to help the audio system engineer properly select and apply transformers. In the interest of simplicity, only basic concepts of their design and manufacture will be discussed.

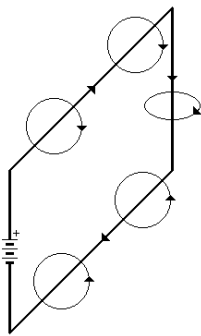## 1.1 Basic Principles and Terminology

1.1.1 Magnetic Fields and Induction

Figure 1 - Magnetic Field Surrounding Conductor

Figure 2 - AC Magnetic Field

As shown in **Figure 1**, a *magnetic field* is created around any conductor (wire) in which current flows. The strength of the field is directly proportional to current. These invisible magnetic *lines of force*, collectively called *flux*, are set up at right angles to the wire and have a direction, or magnetic polarity, which depends on the direction of current flow. Note that although the flux around the upper and lower wires have different directions, the lines inside the loop aid because they point in the same direction. If an alternating current flows in the loop, the instantaneous intensity and polarity of the flux will vary at the same frequency and in direct proportion to the instantaneous current. We can visualize this flux, represented by the concentric circle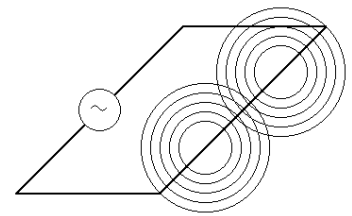s in **Figure 2**, as expanding, contracting, and reversing in polarity with each cycle of the ac current. The law of induction states that a voltage will be *induced* in a conductor exposed to changing flux and that the induced voltage will be proportional to the rate of the flux change. This voltage has an instantaneous polarity which opposes the original current flow in the wire, creating an apparent resistance called *inductive reactance*. Inductive reactance is calculated according to the formula $X_L = 2\delta fL$, where X_L is inductive reactance in ohms, f is frequency in Hz, and L is *inductance* in Henries. An *inductor* generally consists of many turns or loops of wire called a *coil*, as shown in **Figure 3**, which links and concentrates magnetic flux lines, increasing the *flux density*. The inductance of any given coil is determined by factors such as the number of turns, the physical dimensions and nature of the winding, and the properties of materials in the path of the magnetic flux.
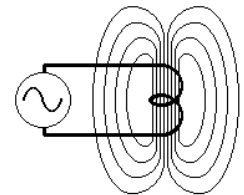

Figure 3 - Coil Concentrates Flux

According to the law of induction, a voltage will be induced in any conductor (wire) that cuts flux lines. Therefore, if we place two coils near each other as shown in **Figure 4**, an ac current in one coil will induce an ac voltage in the second coil. This is the essential principle of energy transfer in a ***transformer***. Because they require a changing magnetic field to operate, transformers will not work at dc. In an ideal transformer, the magnetic coupling between the two coils is total and complete, i.e., all the flux lines from one cut across all the turns of the other. The *coupling coefficient* is said to be unity or 1.00.
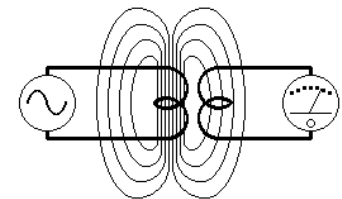
Figure 4 - Inductive Coupling

1.1.2 Windings and Turns Ratio

The coil or winding that is driven by an electrical source is called the *primary* and the other is called the *secondary*. The ratio of the number of turns on the primary to the number of turns on the secondary is called the *turns ratio*. Since essentially the same voltage is induced in each turn of each winding, the primary to secondary *voltage ratio* is the same as the turns ratio. For example, with 100 turns on the primary and 50 turns on the secondary, the turns ratio is 2:1. Therefore, if 20 volts were applied to the primary, 10 volts would appear at the secondary. Since it reduces voltage, this transformer would be called a *step-down* transformer. Conversely, a transformer with a turns ratio of 1:2 would be called a *step-up* transformer since its secondary voltage would be twice that of the primary. Since a transformer does not create power, the power output from the secondary of an ideal transformer can only equal (and in a real transformer only be less than) the power input to the primary. Consider an ideal 1:2 step-up transformer. When 10 volts is applied to its primary, 20 volts appears at its secondary. Since no current is drawn by the

primary, its impedance appears to be infinite or an open circuit. When a 20 Ù load is connected to the secondary, a current of 1 amp flows making output power equal 20 watts. At the same time, a current of 2 amps is drawn by the primary, making input power equal 20 watts. Since the primary is now drawing 2 amps with 10 volts applied, its impedance appears to be 5 Ù. In other words, the 20 Ù load impedance on the secondary has been *reflected* to the primary as 5 Ù. In this example, a transformer with a 1:2 turns ratio exhibited an impedance ratio of 1:4. Transformers always reflect impedances from one winding to another by the square of the their turns ratio or, expressed as a formula: $Zp/Zs = (Np/Ns)^2$ where Zp is primary impedance, Zs is secondary impedance, and Np/Ns is turns ratio (which is the same as the voltage ratio).

The direction in which coils are wound, i.e., clockwise or counter-clockwise, and/or the connections to the start or finish of each winding determines the instantaneous *polarity* of the ac voltages. All windings which are wound in the same direction will have the same polarity between start and finish ends. Therefore, relative to the primary, polarity can be inverted by either (1) winding the primary and secondary in opposite directions, or (2) reversing the start and finish connections to either winding. In schematic symbols for transformers, dots are sometimes used to indicate which ends of windings have the same polarity. Observing polarity is essential when making series or parallel connections to transformers with multiple windings. *Taps* are connections made at any intermediate point in a winding. If 50 turns are wound, an electrical connection brought out, and another 50 turns completes the winding for example, the 100-turn winding is said to be *center-tapped*.
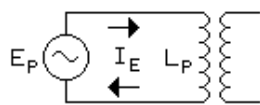
1.1.3 Excitation Current

Figure 5 - Excitation Current

Figure 6 - Cancellation of Flux
Generated by Load Current

As shown in **Figure 5**, when there is no load on the secondary of a transformer and an ac voltage is applied to the primary, an *excitation current* will flow in the primary creating magnetic excitation flux around the winding. In theory, the current is due only to the inductive reactance of the primary winding. I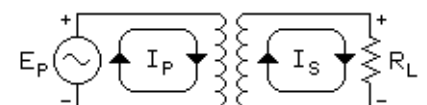n accordance with Ohm's law and the formula for inductive reactance, $I_E = E_P \div 2\eth fL_P$ where I_E is excitation current in amperes, E_P is primary voltage in volts, f is frequency in Hz, and L_P is primary inductance in Henries. In an ideal transformer, primary inductance would be infinite, making excitation current zero. As shown in **Figure 6**, when a load is connected, current will flow in the secondary winding. Because secondary current flow is in the opposite direction, it creates magnetic flux which opposes the excitation flux. This causes the impedance of the primary winding to drop, resulting in additional current being drawn from the driving source, which creates additional flux just sufficient to completely cancel that created by the secondary. The result, which may surprise some, is that flux density in a transformer is not increased by load current. This also illustrates how load current on the secondary is reflected to the primary.

Figure 7 illustrates the relationships between voltage, excitation current, and flux in a transformer as frequency is changed. The horizontal scale is time. The primary voltage Ep is held at a constant voltage as the frequency is tripled and then tripled again. For example, the left waveform could represent one cycle at 100 Hz, the middle 300 Hz, and the right 900 Hz. Because of the primary inductance, excitation current Ip will decrease linearly with frequency, i.e., halving for every doubling in frequency or decreasing at 6 dB per octave. The magnitude of the magnetic flux will likewise *decrease* exactly the same way. Note that the inductance causes a 90-degree phase lag as well. Since the slew rate of a constant amplitude sine wave *increases* linearly with frequency, i.e., doubling for every doubling in frequency or increasing at 6 dB per octave, the resultant flux rate of change remains *constant*. Note that the slope of the Ip and flux waveforms stays constant as frequency is changed. Since, according to the law of induction, the voltage induced in the secondary is proportional to this slope or rate of change, frequency response will be uniform or "flat."
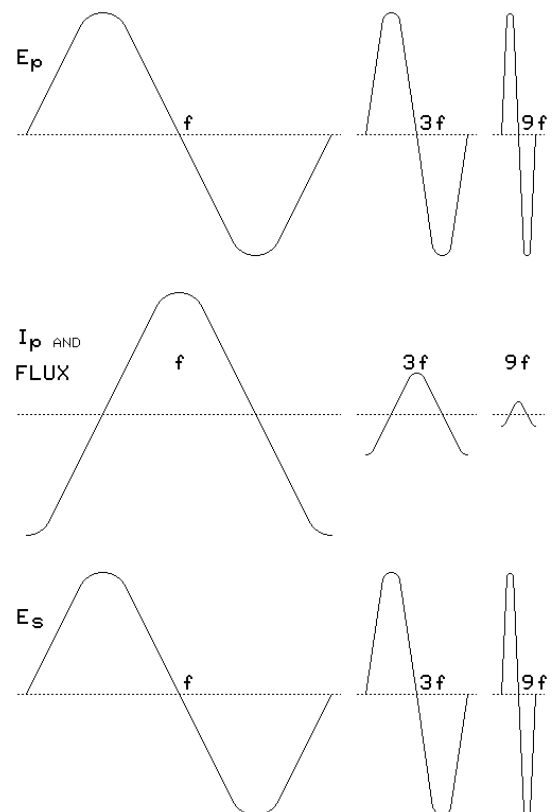

Figure 7 - Excitation Current and Flux
Vary Inversely with Frequency

Audio Transformers
Handbook for Sound Engineers, 3rd Edition

1.2 Realities of Practical Transformers

Thus far, we have not considered the unavoidable parasitic elements which exist in any practical transformer. Even the design of a relatively simple 60 Hz power transformer must take them into account. The design of an audio transformer operating over a 20 Hz to 20 kHz frequency range is much more difficult because these elements often interact in complex ways. For example, materials and techniques which improve low-frequency performance are often detrimental to high-frequency performance and vice-versa. Good transformer designs must consider both the surrounding electronic circuitry and the performance ramifications of internal design tradeoffs.

A schematic representation of the major low-frequency parasitic elements in a generalized transformer is shown in **Figure 8**. The "IDEAL XFMR" represents a perfect transformer having a turns ratio of 1:N and no parasitic elements of any kind. The actual transformer is connected at the "PRI" terminals to the driving voltage source, through its source impedance RG, and at the "SEC" terminals to the load RL.
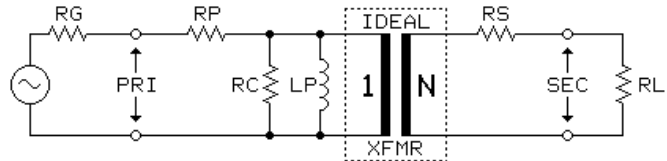

Figure 8 - Transformer Low-Frequency Parasitic Elements

One of the main goals in the design of any transformer is to reduce the excitation current in the primary winding to negligible levels so as not to become a significant load on the driving source. At a given source voltage and frequency, primary excitation current can be reduced only by increasing inductance LP. In the context of normal electronic circuit impedances, very large values of inductance are required for satisfactory operation at the lowest audio frequencies. Of course, inductance can be raised by using a very large number of coil turns but, for reasons discussed later, there are practical limits due to other considerations. Another way to increase inductance by a factor of 10,000 or more is to wind the coil around certain highly magnetic materials.

1.2.1 Core Materials and Construction

Magnetic circuits are quite similar to electric circuits. As shown in **Figure 11**, magnetic flux always takes a closed path from one magnetic pole to the other and, like an electric current, always favors the paths of highest conductivity or least resistance. The equivalent of applied voltage in magnetic circuits is *magnetizing force*, symbolized **H**. It is directly proportional to "ampere-turns" (coil current **I** times its number of turns **N**) and inversely proportional to the flux path length ℓ in the magnetic circuit. The equivalent of electric current flow is *flux density*, symbolized **B**. It is measured as the number of magnetic flux lines per square unit of area. A graphic plot of the relationship between field intensity and flux density is shown in **Figure 9** and is referred to a the "B-H loop" or "hysteresis loop" for a given material. In the United States, the most commonly used units for magnetizing force and flux density are the Oersted and Gauss, respectively, which are CGS (centimeter, gram, second) units. In Europe, the SI (Système International) units amperes per meter and Tesla, respectively, are more common. The slope of the B-H loop indicates how an incremental increase in applied magnetizing force changes the resulting flux density. This slope is effectively a measure of conductivity in the magnetic circuit and is called *permeability*, symbolized **ì**. Any material inside a coil, which can also serve as a form to support it, is called a *core*. By definition, the permeability of air is 1.00 and common "non-magnetic" materials such as aluminum, brass, copper, paper, glass, and plastic also have a permeability of 1 for practical purposes. The permeability of some common "ferro-magnetic" materials is about 300 for ordinary steel, about 5,000 for 4% silicon transformer steel, and up to about 100,000 for some nickel-iron-molybdenum alloys. Because such materials concentrate magnetic flux, they greatly increase the inductance of a coil. Audio transformers must utilize both high-permeability cores and the largest practical number of coil turns to create high primary inductance. Coil inductance increases as the square of the number of turns and in direct proportion to the permeability of the core and can be approximated using the formula: $L = 3.2 \, N^2 \, \text{ì} \, A / 10^8 \, \ell$ where L = inductance in Henries, N = number of coil turns, ì = permeability of core, A = cross-section area of core in square inches, and ℓ = mean flux path length in inches.

The permeability of magnetic materials varies with flux density. As shown in **Figure 9**, when magnetic field intensity becomes high, the material can *saturate*, essentially losing its ability to conduct any additional flux. As a material saturates, its permeability decreases until, at complete saturation, its permeability becomes that of air or 1. In audio transformer applications, magnetic saturation causes low-frequency harmonic distortion to increase steadily for low-frequency signals as they increase in level beyond a threshold. In general, materials with a higher permeability tend to saturate at a lower flux density. In general, permeability also varies inversely with frequency.

Magnetic *hysteresis* can be thought of as a magnetic memory effect. When a magnetizing force saturates material that has high-hysteresis, it remains strongly magnetized even after the force is removed. High-hysteresis materials have wide or "square" B-H loops and are used to make magnetic memory devices and permanent magnets. However, if we magnetically saturate zero-hysteresis material, it will have no residual magnetism (flux density) when the magnetizing force is removed. However, virtually all high-permeability core

Figure 9 - B-H Loop for Magnetic Core Material

materials have some hysteresis, retaining a small memory of their previous magnetic state. Hysteresis can be greatly reduced by using certain metal alloys which have been annealed or heat-treated using special processes. In audio transformers, the non-linearity due to magnetic hysteresis causes increased harmonic distortion for low-frequency signals at relatively low signal levels. Resistor RC in **Figure 8** is a non-linear resistance which represents the combined effects of magnetic saturation, magnetic hysteresis, and eddy-current losses.
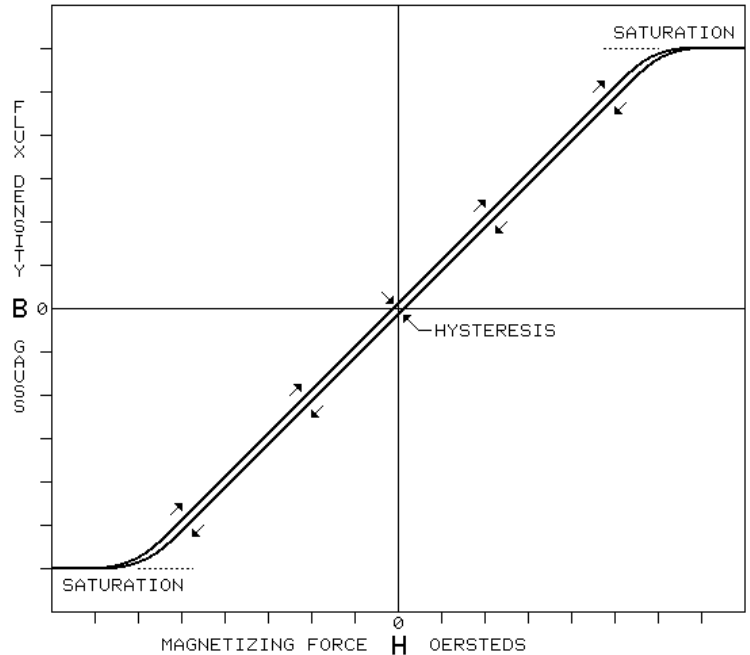
The magnetic *operating point* (or zero signal point) for most transformers is the center of the B-H loop shown in **Figure 9**, where the net magnetizing force is zero. Small ac signals cause a small portion of the loop to be traversed in the direction of the arrows. Large ac signals traverse portions farther from the operating point and may approach the saturation end points. For this normal operating point at the center, signal distortions (discussed in detail later) caused by the curvature of the loop are symmetrical, i.e., they affect the positive excursion and negative excursion equally. Symmetrical distortions produce odd-order harmonics such as third and fifth. If dc current flows in a winding, the operating point will shift to a point on the loop away from the center. This causes the distortion of a superimposed ac signal to become non-symmetrical. Non-symmetrical distortions produce even-order harmonics such as second and fourth. When a small dc current flows in a winding, under say 1% of the saturation value, the effect is to add even-order harmonics to the normal odd-order content of the hysteresis distortion, which affects mostly low-level signals. The same effects occur when the core becomes weakly magnetized, as could happen via the brief accidental application of dc to a winding for example. However, the narrow B-H loop indicates that only a weak residual field would remain even if a magnetizing force strong enough to saturate the core were applied and then removed.

When a larger dc current flows in a winding, the symmetry of saturation distortion is also affected in a similar way. For example, enough dc current might flow in a winding to move the operating point to 50% of the core saturation value. Only half as much ac signal could then be handled before the core would saturate and, when it did, it would occur only for one direction of the signal swing. This would produce strong second-harmonic distortion. To avoid such saturation effects, *air gaps* are sometimes intentionally built into the magnetic circuit. This can be done, for example, by placing a thin paper spacer between the center leg of the E and I cores of **Figure 10**. The magnetic permeability of such a gap is so low — even though it may be only a few thousandths of an inch — compared to the core material, that it effectively controls the flux density in the entire magnetic circuit. Although it drastically reduces the inductance of the coil, gapping is done to prevent flux density from reaching levels which would otherwise saturate the core, especially when substantial dc is present in a winding.
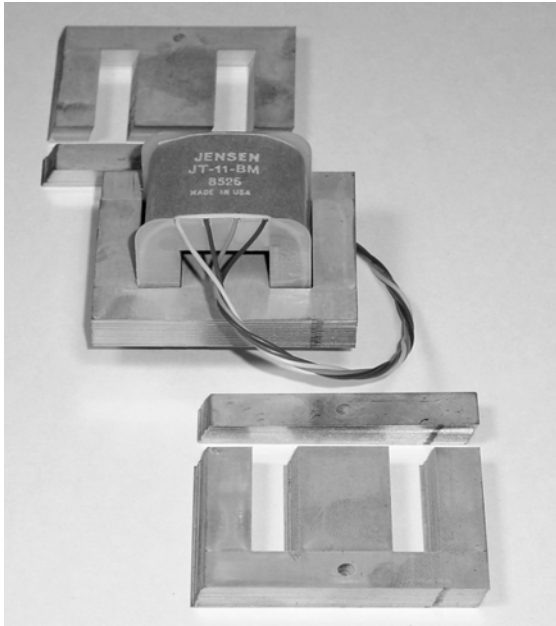
Because high-permeability materials are usually electrical conductors as well, small voltages are also induced in the cross-section of the core material itself giving rise to *eddy currents*. Eddy currents are greatly reduced when the core consists of a "stack" of thin sheets called laminations, as shown in **Figure 10**.

Because the laminations are effectively insulated from each other, eddy currents are generally insignificant. The E and I shaped laminations shown form the widely used "shell" or "double-window" core construction. Its parallel magnetic paths are illustrated in **Figure 11**. When cores are made of laminations, care must be taken that they are flat and straight to avoid tiny air gaps between them which could significantly reduce inductance.
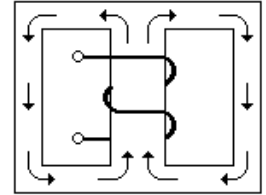

Figure 11 - Magnetic Circuits in Shell Core

Figure 10 - Core Laminations are Stacked and Interleaved around Bobbin which Holds Windings

A *toroidal core* is made by rolling a long thin strip of core material into a coiled ring shape that looks something like a donut. It is insulated with a conformal coating or tape and windings are wound around the core through the center hole using special machines. With a toroidal core, there are no unintended air gaps which can degrade magnetic properties. Audio transformers don't often use toroidal cores because, especially in high-bandwidth designs where multiple sections or Faraday shields are necessary, physical construction becomes very complex. Other core configurations include the *ring core*, sometimes called "semi-toroidal." It is similar to core of **Figure 11** but without the center section and windings are placed on the sides. Sometimes a solid (not laminations) metal version of a ring core is cut into two pieces having polished mating faces. These two *C-cores* are then held together with clamps after the windings are installed.

1.2.2 *Winding Resistances and Auto-Transformers*

If zero-resistance wire existed, some truly amazing transformers could be built. In a 60 Hz power transformer, for example, we could wind a primary with tiny wire on a tiny core to create enough inductance to make excitation current reasonable. Then we could wind a secondary with equally tiny wire. Because the wire has no resistance and the flux density in the core doesn't change with load current, this postage-stamp sized transformer could handle unlimited kilo-watts of power — and it wouldn't even get warm! But, at least until practical superconducting wire is available, real wire has resistance. As primary and secondary currents flow in the winding resistances, the resulting voltage drops cause signal loss in audio transformers and significant heating in power transformers. This resistance can be reduced by using larger (lower gauge) wire or fewer turns, but the required number of turns and the tolerable power loss (or resulting heat) all conspire to force transformers to become physically larger and heavier as their rated power increases. Sometimes silver wire is suggested to replace copper, but since its resistance is only about 6% less, its effect is minimal and certainly not cost-effective. However, there is an alternative configuration of transformer windings, called an *auto-transformer*, which can reduce the size and cost in certain applications. Because an auto-transformer electrically connects primary and secondary windings, it can't be used where electrical isolation is required! In addition, the size and cost advantage is maximum when the required turns ratio is very close to 1:1 and diminishes at higher ratios, becoming minimal in practical designs at about 3:1 or 1:3.

For example, in a hypothetical transformer to convert 100 volts to 140 volts, the primary could have 100 turns and the secondary 140 turns of wire. This transformer, with its 1:1.4 turns ratio, is represented in the upper diagram of **Figure 12**. If 1 amp of secondary (load) current I_S flows, transformer output power is 140 watts and 1.4 amp of primary
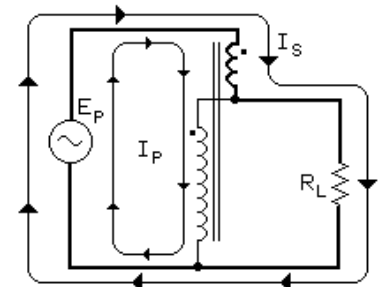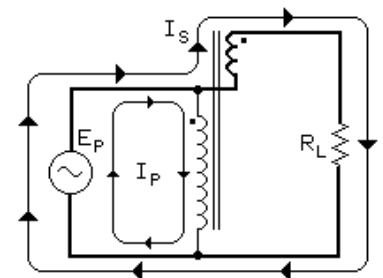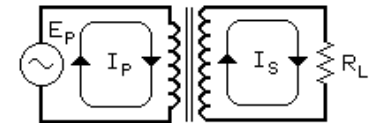

Figure 12 - Auto-Transformers Employ a Buck/Boost Principle

current I_P will flow since input and output power must be equal in the ideal case. In a practical transformer, the wire size for each winding would be chosen to limit voltage losses and/or heating.

An auto-transformer essentially puts the windings in series so that the secondary voltage adds to (boosting) or subtracts from (bucking) the primary input voltage. A step-up auto-transformer is shown in the middle diagram of **Figure 12**. Note that the dots indicate ends of the windings with the same instantaneous polarity. A 40-volt secondary (the upper winding), series connected as shown with the 100-volt primary, would result in an output of 140 volts. Now, if 1 amp of secondary (load) current I_S flows, transformer output power is only 40 watts and only 0.4 amp of primary current I_P will flow. Although the total power delivered to the load is still 140 watts, 100 watts have come directly from the driving source and only 40 watts have been transformed and added by the auto-transformer. In the auto-transformer, 100 turns of smaller wire can be used for the primary and only 40 turns of heavier wire is needed for the secondary. Compare this to the total of 240 turns of heavier wire required in the transformer.

A step-down auto-transformer is shown in the bottom diagram of **Figure 12**. Operation is similar except that the secondary is connected so that its instantaneous polarity subtracts from or bucks the input voltage. For example, we could step down US 120-volt ac power to Japanese 100-volt ac power by configuring a 100-volt to 20-volt step-down transformer as an auto-transformer. Thus, a 100-watt load can be driven using only a 20-watt rated transformer.

The windings of low-level audio transformers may consist of hundreds or even many thousands of turns of wire, sometimes as small as #46 gauge, whose 0.0015 inch diameter is comparable to a human hair. As a result, each winding may have a dc resistance as high as several thousand ohms. Transformer primary and secondary winding resistances are represented by RP and RS, respectively, in **Figure 8**.

1.2.3   *Leakage Inductance and Winding Techniques*

In an ideal transformer, since all flux generated by the primary is linked to the secondary, a short-circuit on the secondary would be reflected to the primary as a short circuit. In real transformers, the unlinked flux causes a residual or *leakage inductance* which can be measured at either winding. Therefore, the secondary would appear to have residual inductance if the primary were shorted and vice-versa. The leakage inductance is shown as LL in the model of **Figure 13**. Note that leakage inductance is

Figure 13 - Transformer High-Frequency Parasitic Elements

reflected from one winding to another as the square of turns ratio, just as other impedances are.

The degree of flux coupling between primary and secondary windings depends on the physical spacing between them and how they are placed with respect to each other. The lowest leakage inductance is achieved by winding the coils on a common axis and as close as possible to each other. The ultimate form of this technique is called *multi-filar* winding where multiple wires are wound simultaneously as if they were a single strand. For example, if two windings (say primary and secondary) are wound as one, the transformer is said to be *bi-filar* wound. Note in the cross-section view of **Figure 14** how the primary and secondary windings are side-by-side throughout the entire winding. Another technique to reduce leakage inductance is to use *layering*, a technique in which portions or *sections* of the primary and/or secondary are wound in sequence over each other to interleave them. For example, **Figure 15** shows the cross-section of a 3-layer transformer where half the primary is wound, then the secondary, followed by the other half of the primary. This results in considerably less leakage inductance than just a secondary over primary 2-layer design. Leakage inductance decreases rapidly as the number of layers is increased.
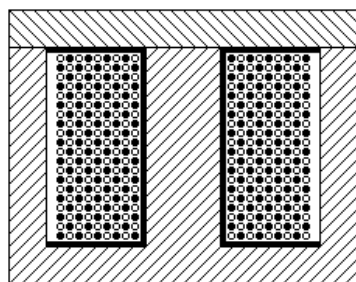
• = PRIMARY   ○ = SECONDARY

Figure 14 - Bi-Filar Windings
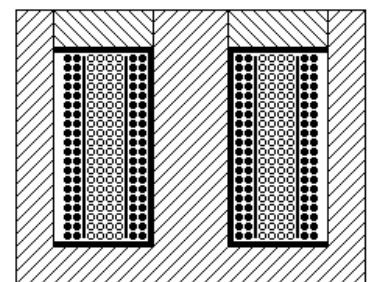
• = PRIMARY   ○ = SECONDARY

Figure 15 - Layered Windings

1.2.4   Winding Capacitances and Faraday Shields

To allow the maximum number of turns in a given space, the insulation on the wire used to wind transformers is very thin. Called "magnet wire," it is most commonly insulated by a thin film of polyurethane enamel. A transformer winding is made, in general, by spinning the bobbin shown in **Figure 10** on a machine similar to a lathe and guiding the wire to form a layer one wire thick across the length of the bobbin. The wire is guided to traverse back and forth across the bobbin to form a coil of many layers as shown in **Figure 15**, where the bobbin cross-section is the solid line on three sides of the winding. This simple side-to-side, back-and-forth winding results in considerable layer-to-layer capacitance within a winding or winding section. More complex techniques such as "universal" winding are sometimes used to substantially reduce winding capacitances. These capacitances within the windings are represented by CP and CS in the circuit model of **Figure 13**. Additional capacitances will exist between the primary and secondary windings and are represented by capacitors CW in the model. Sometimes layers of insulating tape are added to increase the spacing, therefore reducing capacitance, between primary and secondary windings. In the bi-filar windings of **Figure 14**, since the wires of primary and secondary windings are side by side throughout, the inter-winding capacitances CW can be quite high.

In some applications, inter-winding capacitances are very undesirable. Their effects can be almost completely eliminated by the use of a *Faraday shield* between the windings. Sometimes called an *electrostatic shield*, it generally takes the form of a thin sheet of copper foil placed between the windings. Obviously, transformers that utilize multiple layers to reduce leakage inductance will require Faraday shields between all adjacent layers. In **Figure 15** the dark lines between the winding layers are the Faraday shields. Normally, all the shields surrounding a winding are tied together and treated as a single electrical connection. When connected to circuit ground, as shown in **Figure 16**, a Faraday shield intercepts the capacitive current which would otherwise flow between transformer windings.
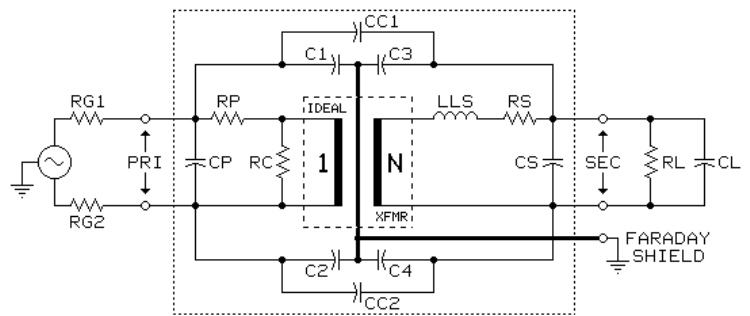

Figure 16 - High-Frequency Equivalent Circuit of Transformer with Faraday Shield and Driven by a Balanced Source

Faraday shields are nearly always used in transformers designed to eliminate "ground noise." In these applications, the transformer is intended to respond only to the voltage difference or signal across its primary and have no response to the noise that exists equally (or common-mode) at the terminals of its primary. A Faraday shield is used to prevent capacitive coupling (via CW in **Figure 13**) of this noise to the secondary. For any winding connected to a balanced line, the matching of capacitances to ground is critical to the rejection of common-mode noise or CMRR, as discussed in Chapter 37. In **Figure 16**, if the primary is driven by a balanced line, C1 and C2 must be very accurately matched to achieve high CMRR. In most applications, such as microphone or line input transformers, the secondary is operated unbalanced, i.e., one side is grounded. This relaxes the matching requirements for capacitances C3 and C4. Although capacitances CC1 and CC2 are generally quite small (a few pF), they have the effect of diminishing CMRR at high audio frequencies and limiting rejection of RF interference.

1.2.5   Magnetic Shielding

A *magnetic shield* has a completely different purpose. Devices such as power transformers, electric motors, and television or computer monitor cathode-ray tubes generate powerful ac magnetic fields. If such a field takes a path through the core of an audio transformer, it can induce an undesired voltage in its windings — most often heard as hum. If the offending source and the victim transformer have fixed locations, orientation of one or both can sometimes nullify the pick-up. In **Figure 11** note that an external field which flows vertically through the core will cause a flux gradient across the length of the coil, inducing a voltage in it, but a field which flows horizontally through the core will not. Such magnetic pick-up is usually worse in "input" transformers (discussed later) because they generally have more turns. It should also be noted that higher permeability core materials are more immune to external fields. Therefore, an unshielded "output" transformer with a high-nickel core will be more immune than one with a steel core.

Another way to prevent such pick-up is to surround the core with a closed (no air gap) magnetic path. This *magnetic shield* most often takes the form of a can or box with tight-fitting lid and is made of high-permeability material. While the permeability of ordinary steel, such as that in electrical conduit, is only about 300, special-purpose nickel alloys can have permeability as high as 100,000. Commercial products include Mumetal®, Permalloy®, HyMu® and Co-Netic®.[1][2] Since the shield completely surrounds the transformer, the offending external field will now flow through it instead of the transformer core. Generally

speaking, care must be taken not to mechanically stress these metals because doing so will significantly decrease their permeability. For this reason, most magnetic shield materials must be re-annealed <u>after</u> they are fabricated.

The effectiveness of magnetic shielding is generally rated in dB. The transformer is placed in an external magnetic field of known strength, generally at 60 Hz. Its output without and with the shield is then compared. For example, a housing of 1/8" thick cast-iron reduces pickup by about 12 dB and a Mumetal can by about 30 dB. Where low-level transformers operate near strong magnetic fields, several progressively smaller shield cans can be nested around the transformer. Two or three Mumetal cans can provide 60 dB and 90 dB of shielding respectively. In very strong fields, because high-permeability materials might saturate, an iron or steel outer can is sometimes used.

Toroidal power transformers can have a weaker radiated magnetic field than other types. Using them can be an advantage if audio transformers must be located near them. However, a toroidal transformer must be otherwise well designed to produce a low external field. For example, every winding must completely cover the full periphery of the core. The attachment points of the transformer lead wires are frequently a problem in this regard. To gain size and cost advantages, most commercial power transformers of any kind are designed to operate on the verge of magnetic saturation of the core. When saturation occurs in any transformer, magnetic field essentially squirts out of the core. Power transformers designed to operate at low flux density will prevent this. Often a standard commercial transformer, when operated at reduced primary voltage, will have a very low external field.

1.3 General Application Considerations

For any given application, a number of parameters must be considered when selecting or designing an appropriate audio transformer. We will discuss how the performance of a transformer can be profoundly affected by its interaction with surrounding circuitry.

1.3.1   Maximum Signal Level, Distortion, and Source Impedance

Because these parameters are inextricably inter-dependent, they must be discussed as a group. Although transformer operating level is often specified in terms of power such as dBm or watts, the only thing that affects distortion is the equivalent driving *voltage*. Distortion is caused by excitation current in the primary winding which is proportional to primary *voltage*, not power. Referring to **Figure 8**, recall that RC represents the distortion producing mechanisms of the core material. Consider that, if both RG (driving source impedance) and RP (internal winding resistance) were zero, the voltage source (by definition, zero impedance) would effectively "short out" RC resulting in zero distortion! But in a real transformer design there is a fixed relationship between signal level, distortion, and source impedance. Since distortion is also a function of magnetic flux density, which increases as frequency decreases, a maximum operating level specification must also specify a frequency. The specified maximum operating level, maximum distortion at a specified low frequency, and maximum allowable source impedance will usually dictate the type of core material which must be used and its physical size. And, of course, cost plays a role, too.

The most commonly used audio transformer core materials are *M6 steel* (a steel alloy containing 6% silicon) and *49% nickel* or *84% nickel* (alloys containing 49% or 84% nickel plus iron and molybdenum). Nickel alloys are substantially more expensive than steel. **Figure 17** shows how the choice of core material affects low-frequency distortion as signal level changes. The increased distortion at low levels is due to magnetic hysteresis and at high levels is due to magnetic saturation. **Figure 18** shows how distortion decreases rapidly with increasing frequency. Because of differences in their hysteresis distortion, the fall-off is most rapid for the 84% nickel and least rapid for the steel. **Figure 19** shows how distortion is strongly affected by the impedance of the driving source (the plots begin at 40 Ù because that is the resistance of the primary winding). Therefore, maximum operating levels predicated on higher frequencies, higher distortion, and lower source impedance will always be higher than those predicated on lower frequencies, lower distortion, and lower source impedance.
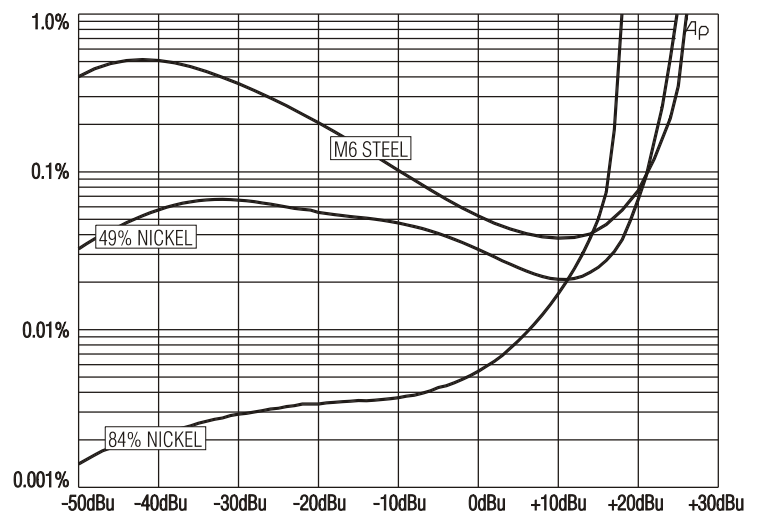

Figure 17 - Measured THD at 20 Hz and 40 Ù Source vs Signal Level for Three Types of Core Material

As background, it should be said that THD or *total* harmonic distortion is a remarkably inadequate way to describe the perceived awfulness of distortion. Distortion consisting of low-order harmonics, 2nd or 3rd for example, is dramatically less audible than that consisting of high-order harmonics, 7th or 13th for example. Consider that, at very low frequencies, even the finest loudspeakers routinely exhibit harmonic distortion in the range of several percent at normal listening levels. Simple distortion tests whose results correlate well with the human auditory experience simply don't exist. Clearly, such perceptions are far too complex to quantify with a single figure.

One type of distortion which is particularly audible is inter-modulation or IM distortion. Tests frequently use a large low-frequency signal and a smaller high-frequency signal and measure how much the amplitude of the high frequency is modulated by the lower frequency. Such inter-modulation creates tones at new, non-harmonic frequencies. The classic SMPTE (Society of Motion Picture and Television Engineers) IM distortion test mixes 60 Hz and 7 kHz signals in a 4:1 amplitude ratio. For virtually all electronic amplifier circuits, there is an approximate relationship between harmonic distortion and SMPTE IM distortion. For example, if an amplifier measured 0.1% THD at 60 Hz at a given operating level, its SMPTE IM distortion would measure about three or four times that, or 0.3% to 0.4% at an equivalent operating level. This correlation is due to the fact that electronic non-linearities generally distort audio signals without regard to frequency. Actually, because of negative feedback and limited gain-bandwidth, most electronic distortions become worse as frequency increases.

Distortion in audio transformers is different in a way which makes it unusually benign. It is caused by the smooth symmetrical curvature of the magnetic transfer characteristic or B-H loop of the core material shown in **Figure 9**. The non-linearity is related to flux density which, for a constant voltage input, is inversely proportional to frequency. The resulting harmonic distortion products are nearly pure third harmonic. In **Figure 18**, note that distortion for 84% nickel

Figure 18 - Measured THD at 0 dBu and 40 Ù Source vs Frequency for the Cores of Figure 16

Figure 19 - Measured THD at 0 dBu and 20 Hz vs Source Impedance for the Cores of Figures 16 and 17

cores roughly quarters for every doubling of frequency, dropping to less than 0.001% above about 50 Hz. Unlike that in amplifiers, the distortion mechanism in a transformer is frequency selective. This makes its IM distortion much less than might be expected. For example, the Jensen JT-10KB-D line input transformer has a THD of about 0.03% for a +26 dBu input at 60 Hz. But, at an equivalent level, its SMPTE IM distortion is only about 0.01% — about a tenth of what it would be for an amplifier having the same THD.

1.3.2   *Frequency Response*

The simplified equivalent circuit of **Figure 20** shows the high-pass RL filter formed by the circuit resistances and transformer primary inductance LP. The effective source impedance is the parallel equivalent of RG + RP and RS + RL. When the inductive reactance of LP equals the effective source impedance, low-frequency response will fall to 3 dB below its mid-band value. For example, consider a transformer having an LP of 10 Henries and winding resistances RP and RS of 50 Ù each. The generator impedance RG is 600 Ù and the load RL is 10 kÙ. The effective source impedance is then (600 Ù + 50 Ù) in parallel with (10 kÙ + 50 Ù) which computes to about 610 Ù. A 10 Henry inductor will have 610 Ù of reactance at about 10 Hz, making response 3 dB down at that frequency. If the generator impedance RG were made 50 Ù instead, response would be ‒3 dB at 1.6 Hz. Lower source impedance will always extend low-frequency bandwidth. Since the filter is single-pole, response falls at 6 dB per octave. As discussed earlier, the permeability of most core material steadily increases as frequency is lowered and typically reaches its
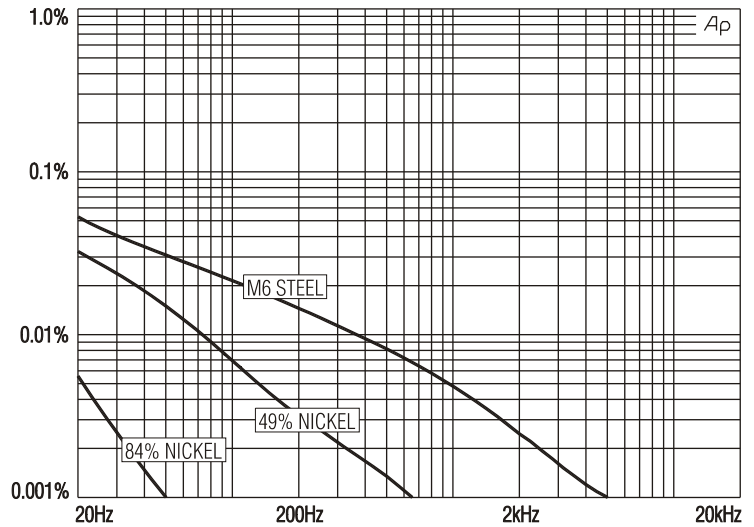
maximum somewhere under 1 Hz. This results in an actual roll-off rate *less* than 6 dB per octave and a corresponding improvement in phase distortion (deviation from linear phase). Although a transformer cannot have response to 0 Hz or dc, it can have much less phase distortion than a coupling capacitor chosen for the same cutoff frequency. Or, as a salesperson might say "it's not a defect, it's a feature."
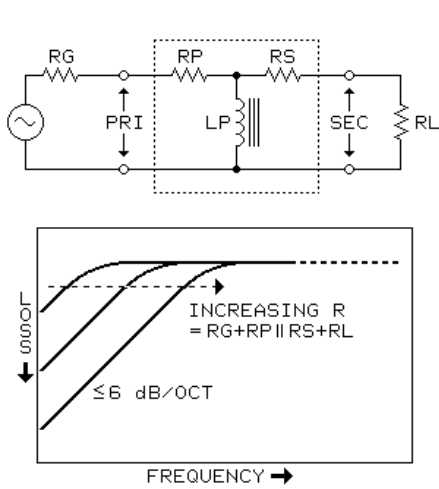

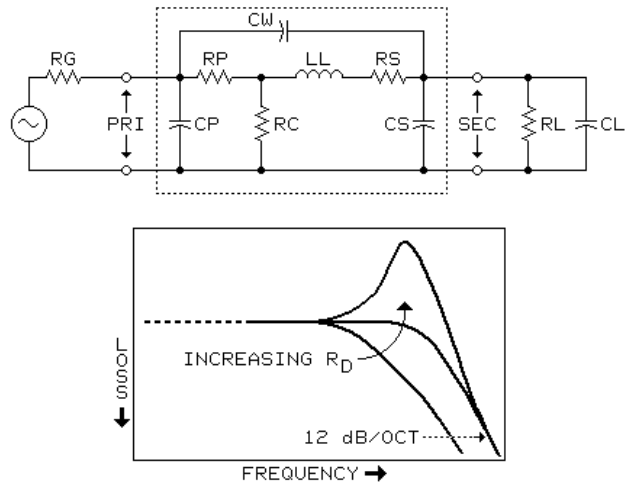**Figure 20** - Simplified Low Frequency Transformer Equivalent Circuit

Figure 21 - Simplified High-Frequency Transformer Equivalent Circuit

The simplified equivalent schematic of **Figure 21** shows the parasitic elements which limit and control high-frequency response. Except in bi-filar wound types discussed below, leakage inductance LL and load capacitance are the major limiting factors. This is especially true when Faraday shields because of the increase in leakage inductance. Note that a low-pass filter is formed by series leakage inductance LL with shunt winding capacitance CS plus external load capacitance CL. Since this filter has two reactive elements, it is a two-pole filter subject to response variations caused by *damping*. Resistive elements in a filter provide damping, dissipating energy when the inductive and capacitive elements resonate. As shown in the figure, if damping resistance R_D is too high, response will rise before it falls and if damping resistance is too low, response falls too early. Optimum damping results in the widest bandwidth with no response peak. It should be noted that placing capacitive loads CL on transformers with high leakage inductance not only lowers their bandwidth but changes the resistance required for optimum damping. For most transformers, RL controls damping. In the time domain, under-damping manifests itself as ringing on square-waves as shown in **Figure 22**. When loaded by its specified load resistance, the same transformer responds as shown in **Figure 23**. In some transformers, source impedance also provides significant damping.
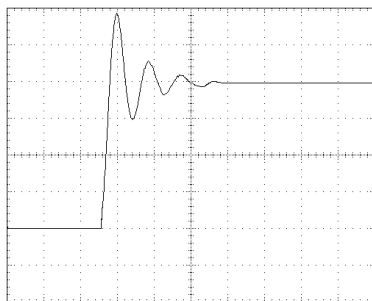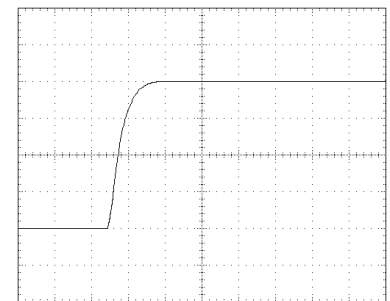

Figure 22 - Undamped Response

Figure 23 - Proper Damping

In bi-filar wound transformers, leakage inductance LL is very low but inter-winding capacitance CW and winding capacitances CP and CS are quite high. Leakage inductance must be kept very small in applications such as line drivers because large cable capacitances CL would otherwise be disastrous to high-frequency response. Also note that a low-pass filter is formed by series RG and shunt CP plus CS. Therefore, driving sources may limit high-frequency response if their source impedance RG is too high. In normal 1:1 bi-filar output transformer designs, CW actually works to *capacitively* couple very high frequencies between windings. Depending on the application, this can be either a defect or a feature.

1.3.3   Insertion Loss

The power output from a transformer will always be slightly less than power input to it. As current flows in its windings, their dc resistance causes additional voltage drops and power loss as heat. Broadly defined, *insertion loss* (or gain) is that caused by inserting a device into the signal path. But, because even an ideal lossless transformer can increase or decrease signal level by virtue of its turns ratio, the term insertion loss is usually defined as the difference in output signal level between the real transformer and an ideal one with the same turns ratio.

The circuit models, Thevenin equivalent circuits, and equations for both ideal and real transformers are shown in **Figure 24**. For example, consider an ideal 1:1 turns ratio transformer and RG = RL = 600 Ù. Since N_s/N_p is 1, the equivalent circuit becomes simply E_i in series with RG or 600 Ù. When RL is connected, a simple voltage divider is formed making $E_o = 0.5\ E_i$ (or a 6.02 dB loss). For a real transformer having RP = RS = 50 Ù, the equivalent circuit becomes E_i in series with RG + RP + RS or 700 Ù.

$$\text{IDEAL } E_o = E_i \cdot \left(\frac{N_s}{N_p}\right) \cdot \frac{RL}{RG \cdot \left(\frac{N_s}{N_p}\right)^2 + RL}$$

$$\text{REAL } E_o = E_i \cdot \left(\frac{N_s}{N_p}\right) \cdot \frac{RL}{(RG+RP) \cdot \left(\frac{N_s}{N_p}\right)^2 + RS + RL}$$

$$\Rightarrow \quad \text{INSERTION LOSS (dB)} = 20 \cdot \log\left(\frac{\text{REAL } E_o}{\text{IDEAL } E_o}\right)$$

Figure 24 - Insertion Loss Compares the Outputs of Real and Ideal Transformers

Now, the output $E_o = 0.462\ E_i$ (or a 6.72 dB loss). Therefore, the insertion loss of the transformer is 0.70 dB.

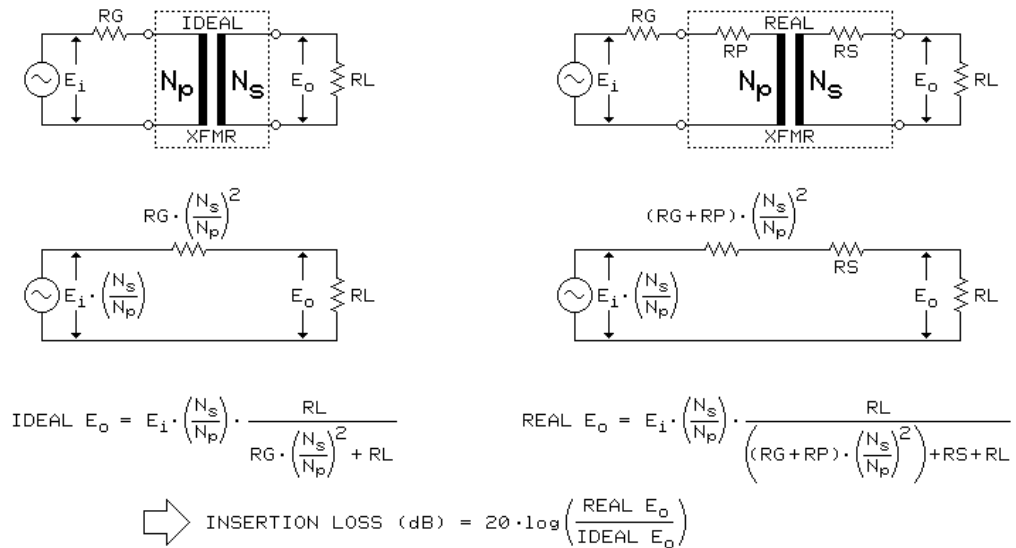Calculations are similar for transformers with turns ratios other than 1:1, except that voltage is multiplied by the turns ratio and reflected impedances are multiplied by the turns ratio squared as shown in the equations. For example, consider a 2:1 turns ratio transformer, RG = 600 Ù, and RL = 150 Ù. The ideal transformer output appears as 0.5 E_i in series with RG/4 or 150 Ù. When RL is connected, a simple voltage divider is formed making $E_o = 0.25\ E_i$ (or a 12.04 dB loss). For a real transformer having RP = 50 Ù and RS = 25 Ù, the equivalent circuit becomes 0.5 E_i in series with (RG + RP)/4 + RS or 187.5 Ù. Now, the output $E_o = 0.222$ E_i (or a 13.07 dB loss). Therefore, the insertion loss of this transformer is 1.03 dB.

1.3.4 *Sources with Zero Impedance*

One effect of using negative feedback around a high-gain amplifier is to reduce output impedance. Output impedance is reduced by the feedback factor which is open-loop gain in dB minus closed-loop gain in dB. A typical op-amp with an open-loop gain of 80 dB, set for closed-loop gain of 20 dB (feedback factor is 80 dB – 20 dB = 60 dB or 1000) will have its open-loop output impedance of 50 Ù reduced by the feedback factor to about 0.05 Ù. Within the limits of linear operation, i.e., no current limiting or voltage clipping, the feedback around the amplifier forces the output to remain constant regardless of loading. For all practical purposes this can be considered a <u>true</u> voltage source.

As seen in **Figure 19**, the distortion performance of ANY transformer is significantly improved when the driving source impedance is less than the dc resistance of the primary. However, little is gained below about 10% of the winding dc resistance. For example, consider a typical line output transformer with a primary dc resistance of 40 Ù. A driving source impedance well under 4 Ù will result in lowest distortion. The line drivers shown in **Figure 28** and **Figure 29** use a paralleled inductor and resistor to isolate or decouple the amplifier from the destabilizing effects of load (cable) capacitance at very high frequencies. Because its impedance is well under an ohm at all audio frequencies, it is much preferred to the relatively large series or "build-out" resistor often used for the purpose. It is even possible for an amplifier to generate *negative* output resistance to cancel the winding resistance of the output transformer. Audio Precision uses such a patented circuit in their System 1 audio generator to reduce transformer-related distortion to extremely low levels.

1.3.5 *Bi-Directional Reflection of Impedances*

The impedances associated with audio transformers seems to confuse many. Much of the confusion probably stems from the fact that *transformers can simultaneously reflect two different impedances*. One is the impedance of the driving source, as seen from the secondary, and the other is the impedance of the load, as seen from the primary. Transformers simply reflect impedances, modified by the square of their turns ratio, from one winding to another. However, because of their internal parasitic elements, transformers tend to produce optimum results when used within a specified range of external impedances.

There is essentially no intrinsic impedance associated with the transformer itself. With no load on its secondary, the primary of a transformer is just an inductor and its impedance will vary linearly with frequency. For example, a 5 H primary winding would

have an input impedance of about 3 kΩ at 100 Hz, 30 kΩ at 1 kHz, and 300 kΩ at 10 kHz. In a proper transformer design, this self-impedance, as well as those of other internal parasitics, should have negligible effects on circuit operation. The following applications will illustrate the point.

A 1:1 *output* transformer application is shown in **Figure 25**. It has a winding inductance of about 25 H and negligible leakage inductance. The open circuit impedance, at 1 kHz, of either winding is about 150 kΩ. Since the DC resistance is about 40 Ω per winding, if the primary is short circuited, the secondary impedance will be 80 Ω. If we place the transformer between a zero-impedance amplifier (more on that later) and a load, the amplifier will "see" the load through the transformer and the load will "see" the amplifier through the transformer. In our example, the amplifier would "look like" 80 Ω to the output line or load and the 600 Ω load would "look like" 680 Ω to the amplifier. If the load were 20 kΩ, it would "look like" slightly less than 20 kΩ because the open circuit transformer impedance (150 kΩ at 1 kHz) is effectively in parallel with it. For most loads, this effect is negligible.
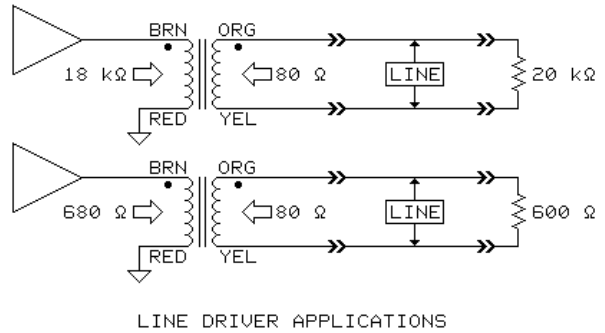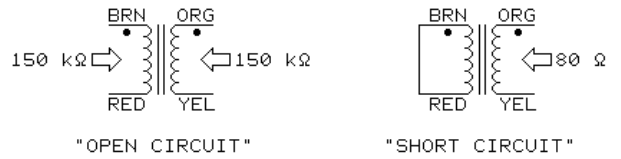
A 4:1 *input* transformer example is shown in **Figure 26**. It has a primary inductance of about 300 H and negligible winding capacitance. The open circuit impedance, at 1 kHz, of the primary is about 2 MΩ. Because this transformer has a 4:1 turns ratio, therefore 16:1 impedance ratio, the secondary open circuit impedance is about 125 kΩ. The DC resistances are about 2.5 kΩ for the primary and 92 Ω for the secondary. Since this is an input transformer, it must be used with the specified secondary load resistance of 2.43 kΩ for proper damping (flat frequency response). This load on the secondary will be transformed by the turns ratio to "look like" about 42 kΩ at the primary. To minimize the noise contribution of the amplifier stage, we need to know what the transformer secondary "looks like," impedance wise, to the amplifier. If we assume that the primary is driven from the line in our previous output transformer example with its 80 Ω source impedance, we can calculate that the secondary will "look like" about 225 Ω to the amplifier input. Actually, any source impedance less than 1 kΩ would have little effect on the impedance seen at the secondary.
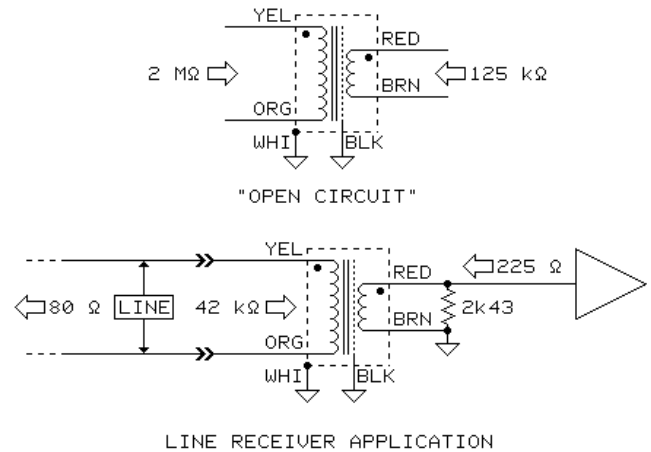

Figure 25 - Impedance Reflection in a 1:1 Transformer

Figure 26 - Impedance Reflection in a 4:1 Transformer

Transformers are not "intelligent" — they can't isolate, in the loading sense, outputs from one another or magically couple signals in one direction only. Magnetic coupling is truly bi-directional. For example, **Figure 27** shows a three-winding 1:1:1 transformer connected to drive two 600 Ω loads. The driver "sees" the loads in parallel or, neglecting winding resistances, 300 Ω. Likewise, a short on either output will be reflected to the driver as a short. Of course, turns ratios and winding resistances must be taken into account to calculate actual driver loading. For the same reason, stereo L and R outputs driving two windings on the same transformer are effectively driving each other, possibly causing distortion or damage.
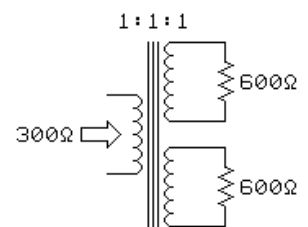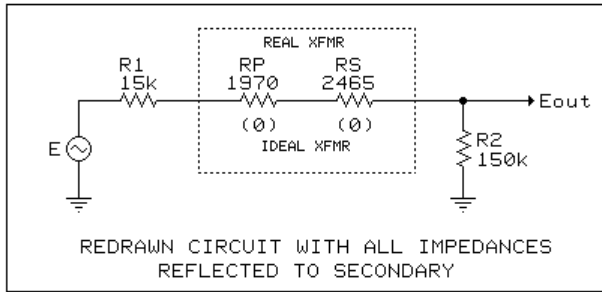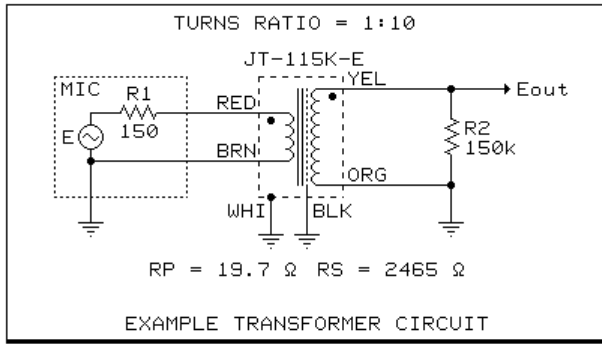
1.3.6 *Transformer Noise Figure*

Figure 27 - Multiple Loads are Effectively Paralleled

Although the step-up turns ratio of a transformer may provide "noise-free" voltage gain, some 20 dB for a 1:10 turns ratio, it's important to understand that improvements in signal-to-noise ratio are not solely due to this gain. Because most amplifying devices generate current noise as well as voltage noise, their noise performance will suffer when turns ratio is above the optimum (see Chapter 21 on mic preamps). *Noise figure* measures, in dB, how much the output signal-to-noise ratio of a system is degraded by a given system component. All resistances, including the winding resistances of transformers, generate thermal noise. Therefore, the noise figure of a transformer indicates the increase in thermal noise or hiss when it replaces an ideal noiseless transformer having the same turns ratio, i.e., voltage gain. The noise figure of a transformer is calculated as follows:

TURNS RATIO = 1:10

JT-115K-E

MIC · R1 · RED · YEL · Eout
E · 150 · BRN · R2 150k
WHI · BLK · ORG

RP = 19.7 Ω   RS = 2465 Ω

EXAMPLE TRANSFORMER CIRCUIT

REAL XFMR
R1 15k · RP 1970 · RS 2465 · Eout
(0) (0)
IDEAL XFMR
E · R2 150k

REDRAWN CIRCUIT WITH ALL IMPEDANCES
REFLECTED TO SECONDARY

PRIMARY SIDE IMPEDANCES GET MULTIPLIED
BY THE SQUARE OF THE TURNS RATIO.

1) R1 = 150 Ohms × (10)^2 = 15k

2) RP = 19.7 Ohms × (10)^2 = 1970

3) RS = SECONDARY DCR = 2465

4) R2 = 150k (LOAD)

THE TRANSFORMER NOISE FIGURE IS CALCULATED BY
COMPARING A REAL TRANSFORMER WITH ITS WINDING
RESISTANCES TO AN IDEAL TRANSFORMER WITH NO
WINDING RESISTANCES.

FIRST, TRANSFORM ALL IMPEDANCES TO THE SECONDARY
AS SHOWN TO THE LEFT.

THERE ARE TWO COMPONENTS TO THE CALCULATION:

1) THE ADDITIONAL NOISE DUE TO THE
   INCREASED OUTPUT IMPEDANCE.

$$\text{REAL Zout} \quad \frac{150k \times (15k + 1970 + 2465)}{150k + 15k + 1970 + 2465} = 17k206$$

$$\text{IDEAL Zout} \quad \frac{150k \times 15k}{150k + 15k} = 13k636$$

$$NF = 20 \, Log \sqrt{\frac{17k206 \, (REAL)}{13k636 \, (IDEAL)}} = 1.01dB$$

2) THE DECREASE IN SIGNAL LEVEL AT THE OUTPUT
   DUE TO THE INCREASED SERIES LOSSES.

$$\text{IDEAL Eout} \quad \frac{150k}{150k + 15k} = 0.909$$

$$\text{REAL Eout} \quad \frac{150k}{150k + 15k + 1970 + 2465} = 0.885$$

$$NF = 20 \, Log \, \frac{0.909 \, (IDEAL)}{0.885 \, (REAL)} = 0.232dB$$

3) TOTAL NF = 1.01dB + 0.232dB = 1.233dB

1.3.7   *Basic Classification by Application*

Many aspects of transformer performance, such as level-handling, distortion, and bandwidth, depend critically on the impedance of the driving source and, in some cases, the resistance and capacitance of the load. These impedances play such an important role that they essentially classify audio transformers into two basic types. Most simply stated, *output* transformers are used when load impedances are low, as in line drivers, while *input* transformers are used when load impedances are high, as in line receivers. The conflicting technical requirements for output and input types make their design and physical construction very different. Of course, some audio transformer applications need features of both input and output transformers and are not so easily classified.

Output transformers must have very low leakage inductance in order to maintain high-frequency bandwidth with capacitive loads. Because of this, they rarely use Faraday shields and are often multi-filar wound. For low insertion loss, they use relatively few turns of large wire to decrease winding resistances. Since they use fewer turns and operate at relatively high signal levels, output transformers seldom use magnetic shielding. On the other hand, input transformers directly drive the usually high-resistance, low-capacitance input of amplifier circuitry. Many input transformers operate at relatively low signal levels, frequently have a Faraday shield, and are usually enclosed in at least one magnetic shield.

2   Audio Transformers for Specific Applications

Broadly speaking, audio transformers are used because they have two very useful properties. First, they can benefit circuit performance by transforming circuit impedances, to optimize amplifier noise performance for example. Second, because there is no direct electrical connection between its primary and secondary windings, a transformer provides electrical or galvanic *isolation* between two circuits. As discussed in Chapter 37, isolation in signal circuits is a powerful technique to prevent or cure noise problems caused by normal ground voltage differences in audio systems. To be truly useful, a transformer should take full advantage of one or both of these properties but not compromise audio performance in terms of bandwidth, distortion, or noise.

2.1 Equipment-Level Applications

2.1.1 *Microphone Input*

A microphone input transformer is driven by the nominal 150 Ù (or 200 Ù in Europe) source impedance of professional microphones. One of its most important functions is to transform this impedance to a generally higher one more suited to optimum noise performance. As discussed in chapter 21, this optimum impedance may range from 500 Ù to over 15 kÙ, depending on the amplifier. For this reason, microphone input transformers are made with turns ratios ranging from 1:2 to 1:10 or higher. The

Figure 28 - Microphone Pre-Amplifier with 40 dB Overall Gain

circuit of **Figure 28** uses a 1:5 turns ratio transformer, making the microphone appear as a 3.7 kÙ driving source to the IC amplifier, which optimizes its noise. The input impedance of the transformer is about 1.5 kÙ. It is important that this impedance remain reasonably flat with frequency to avoid altering microphone response by loading it excessively at frequency extremes.
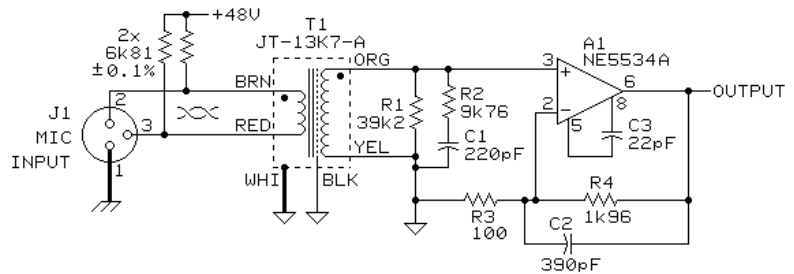
In all balanced signal connections, common-mode noise can exist due to ground voltage differences or magnetic or electrostatic fields acting on the inter-connecting cable. It is called common-mode noise because it appears equally on the two signal lines, at least in theory. Perhaps the most important function of a balanced input is to reject (not respond to) this common-mode noise. A figure comparing the ratio of its differential or normal signal response to its common-mode response is called common-mode rejection ratio or CMRR. An input transformer must have two attributes to achieve high CMRR. First, the capacitances of its two inputs (to ground) must be very well matched and as low as possible. Second, it must have minimal capacitance between its primary and secondary windings. This is usually accomplished by precision winding of the primary to evenly distribute capacitances and the incorporation of a Faraday shield between primary and secondary. Because the common-mode input impedances of a transformer consist only of capacitances of about 50 pF, transformer CMRR is maintained in real-world systems where the source impedances of devices driving the balanced line and the capacitances of the cable itself are not matched with great precision [3].

Because tolerable common-mode voltage is limited only by winding insulation, transformers are well suited for phantom power applications. The standard arrangement using precision resistors is shown in **Figure 28**. Resistors of lesser precision may degrade CMRR. Feeding phantom power through a center tap on the primary requires that both the number of turns and the dc resistance on either side of the tap be precisely matched to avoid small dc offset voltages across the primary. Normal tolerances on winding radius and wire resistance make this a less precise method than the resistor pair in most practical transformer designs. Virtually all microphone input transformers will require loading on the secondary to control high-frequency response. For the circuit in the figure, network R1, R2, and C1 shape the high-frequency response to a Bessel roll-off curve. Because they operate at very low signal levels, most microphone input transformers also have magnetic shielding.

2.1.2 *Line Input*

A line input transformer is driven by a balanced line and, most often, drives a ground-referenced (unbalanced) amplifier stage. As discussed in Chapter 37, modern voltage-matched interconnections require that line inputs have impedances of 10 kÙ or more, traditionally called "bridging." In the circuit of **Figure 29**, a 4:1 step-down transformer is used which has an input impedance of about 40 kÙ.

High common-mode noise rejection or CMRR is achieved in line input transformers using the same techniques as those for microphones. Again, because its common-mode input impedances consist of small capacitances, a good input transformer will exhibit high CMRR even when signal sources

Figure 29 - Low-Noise Unity-Gain Balanced Line Input Stage

are real-world equipment. Electronically-balanced stages, especially simple differential amplifiers, are very susceptible to tiny impedance imbalances in driving sources. However, they usually have impressive CMRR figures when the signal source is a laboratory generator. The pitfalls of measurement techniques will be discussed in section 3.1.
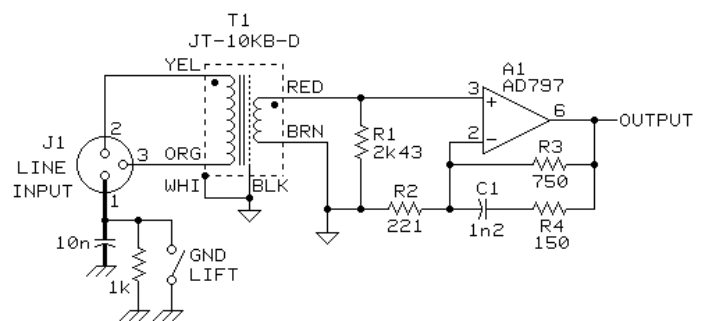
As with any transformer having a Faraday shield, line input transformers have significant leakage inductance and their secondary load effectively controls their frequency response and roll-off characteristics. The load resistance or network recommended by the manufacturer should be used to achieve specified bandwidth and transient response. Input transformers are intended to immediately precede an amplifier stage with minimal input capacitance. Additional capacitive loading of the secondary should be avoided because of its adverse effect on frequency and phase response. For example, the capacitance of two feet of ordinary shielded cable, about 100 pF, is enough to significantly degrade performance of many input transformers.

2.1.3   *Moving-Coil Phono Input*

Moving-coil phonograph pickups are very low-impedance, very low-output devices. Some of them have source impedances as low as 3 Ù, making it nearly impossible to achieve optimum noise performance in an amplifier. The transformer shown in **Figure 30** has a three-section primary that can be series-connected as a 1:4 step-up for 25 Ù to 40 Ù devices and parallel-connected as a 1:12 step-up for 3 Ù to 5 Ù devices. In either case, the amplifier sees a 600 Ù source impedance that enables low-noise operation. The transformer is packaged in double magnetic shield cans and has a Faraday shield. The loading network R1, R2, and C1 tailor the high-frequency response to a Bessel curve.
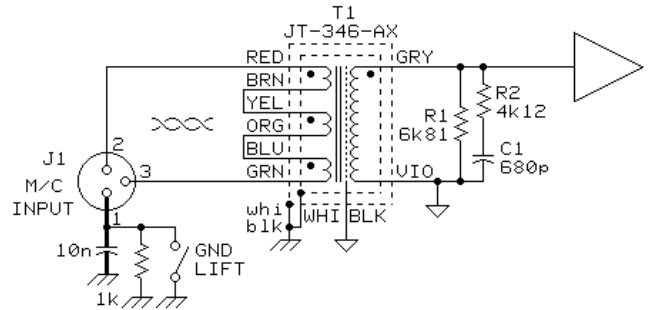

Figure 30 - Preamp for 25 Ù to 40 Ù Moving-Coil Pickups

2.1.4   *Line Output*

A line-level output transformer is driven by an amplifier and typically loaded by several thousand pF of cable capacitance plus the 20 kÙ input impedance of a balanced "bridging" line receiver. At high frequencies, most driver output current is actually used driving the cable capacitance. Sometimes, terminated 150 Ù or 600 Ù lines must be driven, requiring even more driver output current. Therefore, a line output transformer must have a low output impedance that stays low at high frequencies. This requires both low resistance windings and very low leakage inductance, since they are effectively in series between amplifier and load. To maintain impedance balance of the output line, both driving impedances and inter-winding capacitances must be well matched at each end of the windings. A typical bifilar-wound design has winding resistances of 40 Ù each, leakage inductance of a few micro-henries, and a total inter-winding capacitance of about 20 nF matched to within 2% across the windings.
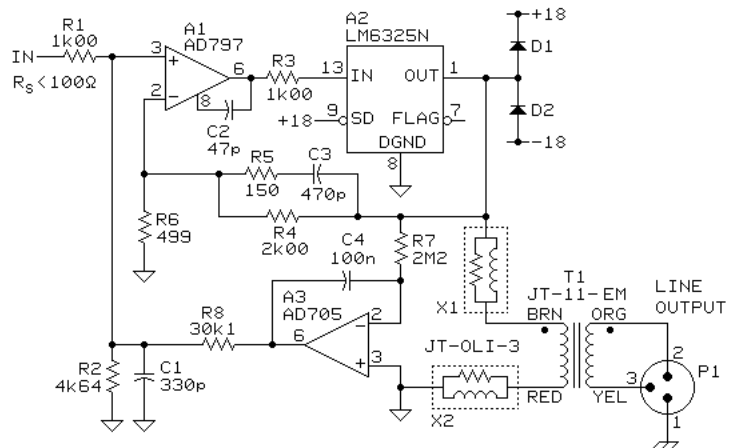

Figure 31 - Typical Line Output Application Circuit

The high-performance circuit of **Figure 31** uses op-amp A1 and current booster A2 in a feedback loop setting overall gain at 12 dB. A3 provides the high gain for a dc servo feedback loop used to keep dc offset at the output of A2 under 100 ì V. This prevents any significant dc flow in the primary of transformer T1. X1 provides capacitive load isolation for the amplifier and X2 serves as a tracking impedance to maintain high-frequency impedance balance of the output. High-conductance diodes D1 and D2 clamp inductive kick to protect A2 in case an unloaded output is driven into hard clipping.

The circuit of **Figure 32** is well suited to the lower signal levels generally used in consumer systems. Because its output floats, it can drive either balanced or unbalanced outputs, but not at the same time. Floating the unbalanced output avoids ground loop problems that are inherent to unbalanced interconnections.

In both previous circuits, because the primary drive of T1 is single-ended, the voltages at the secondary will *not* be symmetrical, especially at high frequencies. THIS IS NOT A
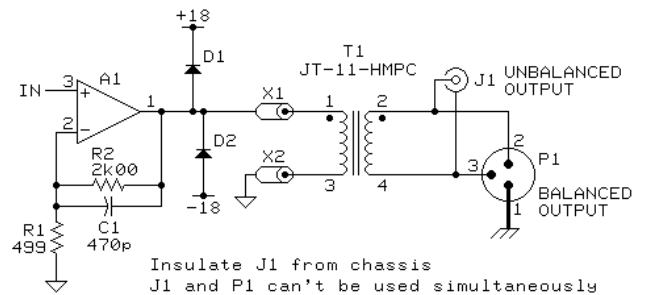

Figure 32 - Universal Isolated Output Application

PROBLEM. Contrary to widespread myth and as explained in Chapter 37, signal symmetry has <u>absolutely nothing</u> to do with noise rejection in a balanced interface! Signal symmetry in this, or any other floating output, will depend on the magnitude and matching of cable and load impedances to ground. If there is a requirement for signal symmetry, the transformer should be driven by dual, phase-inverted drivers.

The circuit of **Figure 33** uses a cathode follower circuit which replaces the usual resistor load in the cathode with an active current sink. The circuit operates at quiescent plate currents of about 10 mA and drives the transformer with a source impedance of about 60 Ù, which is less than 10% of its primary dc resistance. C2 is used to prevent dc flow in the primary. Since the transformer has a 4:1 turns ratio (or 16:1 impedance ratio), a 600 Ù output load is reflected to the driver circuit as about 10 kÙ. Since the signal swings on the primary are four times as large as those on the secondary, high-frequency capacitive coupling is prevented by a Faraday shield. The secondary windings may be parallel connected to drive a 150Ù load. Because of the Faraday shield, output winding capacitances are low and the output signal symmetry will be determined largely by the balance of line and load impedances.
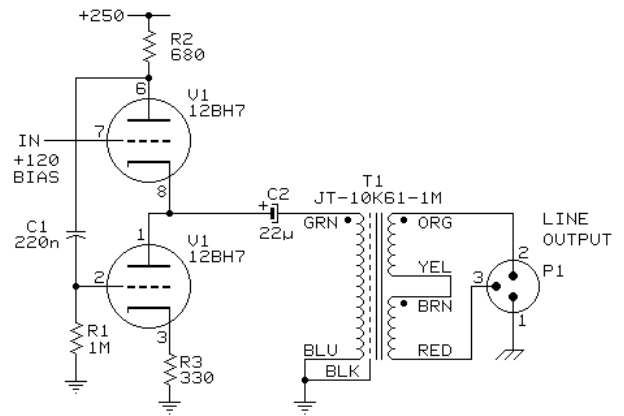
Figure 33 - Double Cathode-Follower Line Driver

2.1.5   *Inter-Stage and Power Output*

Inter-stage coupling transformers are seldom seen in contemporary equipment but were once quite popular in vacuum-tube amplifier designs. They typically use turns ratios in the 1:1 to 1:3 range and, as shown in **Figure 34**, may use a center-tapped secondary producing phase-inverted signals to drive a push-pull output stage. Because both plate and grid circuits are relatively high impedance, windings are sometimes section-wound to reduce capacitances. Resistive loading of the secondary is usually necessary both to provide damping and to present a uniform load impedance to the driving stage. Although uncommon, inter-stage transformers for solid-state circuitry are frequently bifilar wound units similar to line output designs.

The classic push-pull *power output* stage, with many variations over the years, has been used in hi-fi gear, PA systems, and guitar amplifiers. The turns ratio of the output transformer is generally chosen for a reflected load at the tubes of several thousand ohms plate-to-plate. A typical 30:1 turns ratio may require many interleaved sections to achieve bandwidth extending well beyond 20 kHz.
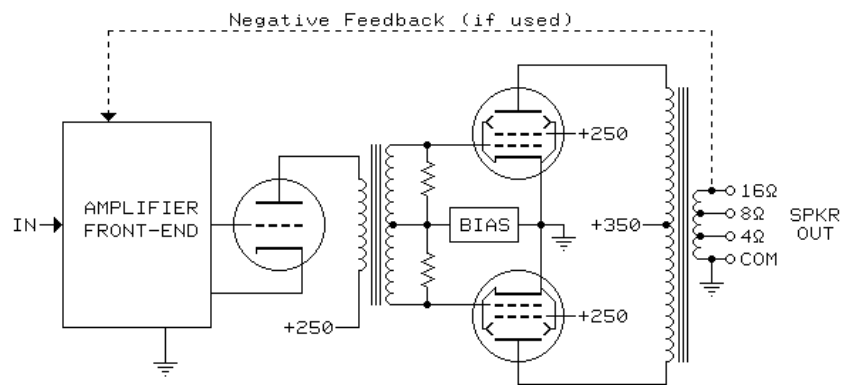
If the quiescent plate currents and the number of turns in each half of the primary winding are matched, magnetic flux in the core will cancel at dc. Since any current-balancing is temporary at best, these transformers nearly always use steel cores. The relatively high driving impedance of

Figure 34 - Push-Pull Vacuum-Tube Power Amplifier

the tube plates results in considerable transformer related distortion. To reduce distortion, feedback around the transformer is often employed. But to achieve stability (freedom from oscillation), very wide bandwidth (actually low phase shift) is required of the transformer when a feedback loop is closed around it. As a result, some of these output transformer designs are very sophisticated. Some legendary wisdom suggests "as a rough guide" that a good-fidelity output transformer should have a core weight and volume of at least 0.34 pounds and 1.4 cubic inches respectively per watt of rated power [4].

A "single-ended" power amplifier is created by removing the lower tube and the lower half of the transformer primary from the circuit of **Figure 34**. Now plate current will create a strong dc field in the core. As discussed in section 1.2.1, the core will likely require an air gap to avoid saturation. This reduces inductance (limiting low-frequency response) and increases even-order distortion products. Such a single-ended pentode power amplifier was widely used in 5-tube table radios of the fifties and sixties.

### 2.1.6	Microphone Output

There are two basic types of output transformers used in microphones, step-up and step-down. In a ribbon microphone, the ribbon element may have an impedance of well under 1 Ù, requiring a step-up transformer with a turns ratio of 1:12 or more to make its nominal output impedance around 150 Ù. Typical dynamic elements have impedances from 10 Ù to 30 Ù, which require step-up turns ratios from 1:2 to 1:4. These step-up designs are similar to line output transformers in that they have no Faraday or magnetic shields, but are smaller because of lower signal levels.

A condenser microphone has integral circuitry to buffer and/or amplify the signal from its extremely high-impedance transducer. Since this low-power circuitry operates from the "phantom" supply, it may be unable to directly drive the 1.5 kÙ

Figure 35 - Condenser Microphone Output Transformer

input impedance of a typical microphone preamp. The output transformer shown in **Figure 35**, which has an 8:1 step-down ratio, will increase the impedance seen by Q1 to about 100 kÙ. Due to its high turns ratio, a Faraday shield is used to prevent capacitive coupling of primary signal to the output.
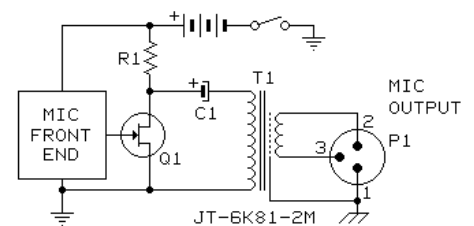
2.2 System-Level Applications

### 2.2.1	Microphone Isolation or "Splitter"

The primary of a transformer with a 1:1 turns ratio can "bridge" the output of a 150 Ù to 200 Ù microphone feeding one pre-amp and the secondary of the transformer can feed a duplicate of the microphone signal to another pre-amp. Of course, a simple "Y" cable could do this but there are potential problems. There are often large and noisy voltages between the grounds of two pre-amplifiers. The isolation provided by the transformer prevents the noise from coupling to the balanced signal line. To reduce capacitive noise coupling, Faraday shields are included in better designs

Figure 36 - A 3-Way Microphone "Splitter" Box

and double Faraday shields in the best. As discussed in section 11.1.3.5, the input impedances of all the pre-amps, as well as all the cable capacitances, will be seen in parallel by the microphone. This places a practical upper limit on how many "ways" the signal can be split. Transformers are commercially available in 2, 3, and 4-winding versions. A 3-way splitter box schematic is shown in **Figure 36**. Since the microphone is *directly* connected only to the "direct" output, it is the only one that can pass phantom power to the microphone. To each preamp, each isolated output "looks like" a normal floating (ungrounded) microphone. The ground lift switches are normally left open to prevent possible high ground current flow in the cable shields.
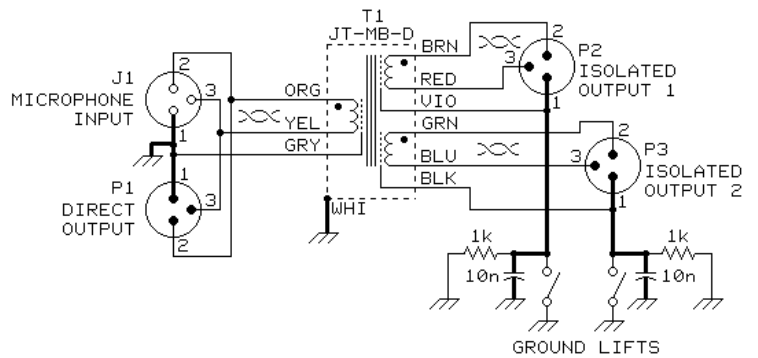
### 2.2.2	Microphone Impedance Conversion

There are some legacy dynamic microphones which are high-impedance (about 50 kÙ) and have two-conductor cable and connector (unbalanced). When such a microphone must be connected to a standard balanced low-impedance microphone pre-amp, a transformer with a turns ratio of about 15:1 is necessary. Similar transformers can be used to adapt a low-impedance microphone to the unbalanced high-impedance input of a legacy pre-amplifier.

### 2.2.3	Line to Microphone Input or "Direct Box"

Because its high-impedance input accepts line-level signals and its output drives the low-level, low-impedance microphone input of a mixing console, the device shown in **Figure 37** is called a "direct box." It is most often driven by an electric guitar, synthesizer, or other stage instrument. Because it uses a transformer, it provides ground isolation as well. In this typical circuit, since the transformer has a 12:1 turns ratio, the impedance ratio is 144:1. When the microphone input has a typical 1.5 kÙ input impedance, the input impedance of the direct box is about 200 kÙ. The transformer shown has a separate Faraday shield for each winding to minimize capacitively coupled ground noise.
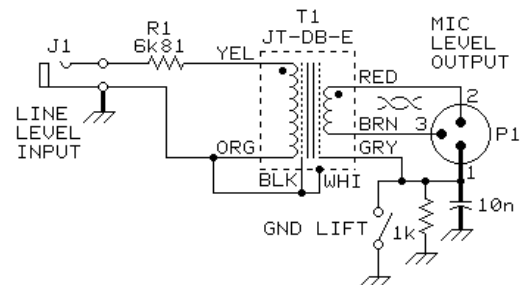
Figure 37 - A Transformer-Isolated "Direct Box"

2.2.4   *Line Isolation or "Hum Eliminators"*

There are a remarkable number of "black boxes" on the market intended to solve "ground loop" problems. This includes quite a number of transformer-based boxes. With rare exception, those boxes contain *output* transformers. Tests were performed to compare noise rejection of the original interface to one with an added output transformer and to one with an added input transformer. The tests accurately simulated typical real-world equipment (see the definitions at the end of this section).

Figure 38 shows results of CMRR tests on a ***balanced*** interface using the IEC 60268-3 test procedure (discussed in section 3.1.2). This test recognizes that the impedances of real-world balanced outputs are not matched with the precision of laboratory equipment. While the *output* transformer reduces 60 Hz "hum" by over 20 dB, it has little effect on "buzz" artifacts over about 1 kHz. The *input* transformer increases rejection to over 120 dB at 60 Hz and to almost 90 dB at 3 kHz, where the human ear is most sensitive to faint sounds.

Figure 39 shows results of ground noise rejection tests on an ***unbalanced*** interface. By definition, there is 0 dB of inherent rejection in an unbalanced interface (see Chapter 37).While the ***output*** transformer reduces 60 Hz "hum" by about 70 dB, it reduces "buzz" artifacts around 3 kHz by only 35 dB. The ***input*** transformer increases rejection to over 100 dB at 60 Hz and to over 65 dB at 3 kHz.

Figure 40 shows results of CMRR tests when an ***unbalanced*** output drives a ***balanced*** input. A two-wire connection of this interface will result in zero rejection (see Chapter 37). Assuming a three-wire connection, the −30 dB plot shows how CMRR of typical electronically-balanced input stages is degraded by the 600 Ù source imbalance. Again, the *output* transformer improves 60 Hz "hum" by over 20 dB, it has little effect on "buzz" artifacts over about 1 kHz. The *input* transformer increases rejection to over 120 dB at 60 Hz and to almost 90 dB at 3 kHz.

Figure 41 shows results of ground noise rejection tests when a ***balanced*** output drives an ***unbalanced*** input. Because our balanced output does not float, the direct connection becomes an unbalanced interface having, by definition, 0 dB of rejection. While the ***output*** transformer reduces 60 Hz "hum" by about 50 dB, it reduces "buzz" artifacts around 3 kHz by less than 20 dB. The ***input*** transformer increases rejection to over 105 dB at 60 Hz and to almost 75 dB at 3 kHz. In this application it is usually desirable to attenuate the signal by about 12 dB (from +4 dBu or 1.228 volts to −10 dBV or 0.316 volts) as well as provide ground isolation. This can be conveniently done by using a 4:1 step-down input transformer such as the one in **Figure 29**, which will produce rejection comparable to that shown here.

One might fairly ask "Why not use a 1:4 step-up transformer when an ***unbalanced*** output drives a ***balanced*** input to get 12 dB of signal gain?" Because of the circuit impedances involved, the answer is because it doesn't work very well. Recall that a 1:4 turns ratio has an impedance ratio of 1:16. This means that the
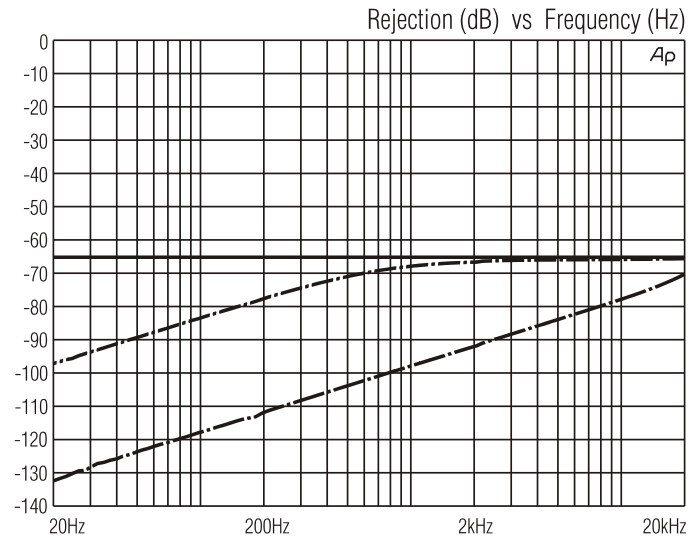

Figure 38 - Balanced Output to Balanced Input
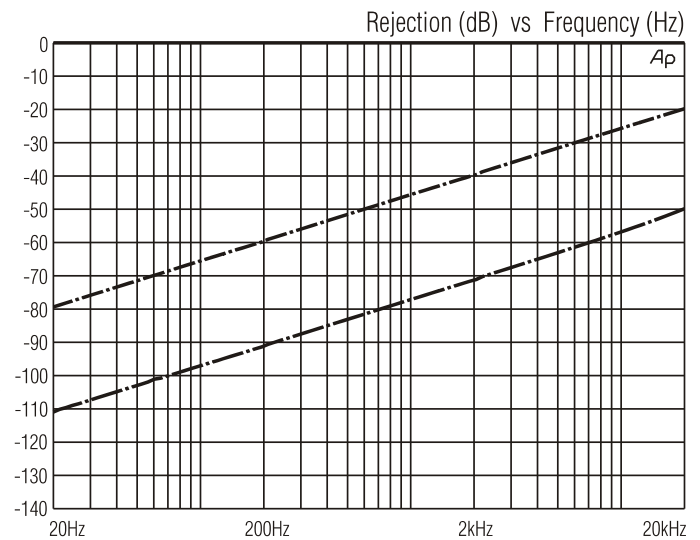Top = None, Center = Output, Bottom = Input

Figure 39 - Unbalanced Output to Unbalanced Input
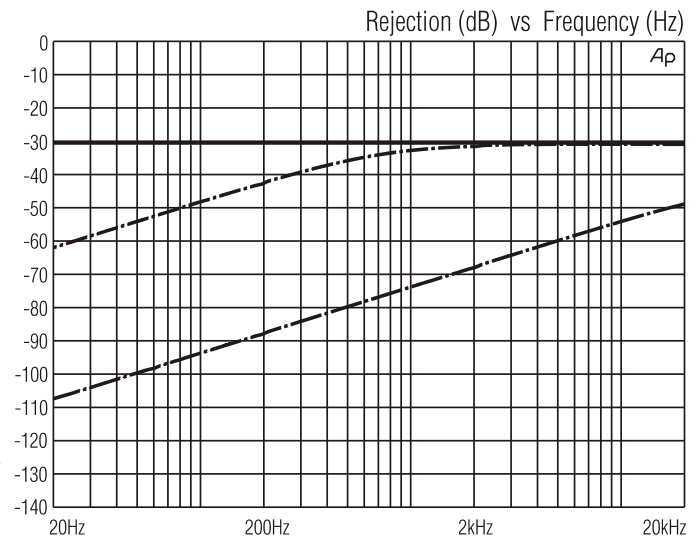Top = None, Center = Output, Bottom = Input

Figure 40 - Unbalanced Output to Balanced Input
Top = None, Center = Output, Bottom = Input

input impedance of the "pro" balanced input we drive will be reflected back to the "consumer" output at one-sixteenth that. Since the source impedance (usually unspecified, but not the same as *load* impedance) of a consumer outputs is commonly 1 kÙ or more, the reflected loading losses are high. A 1:4 step-up transformer would have its own insertion losses, which we will rather optimistically assume at 1 dB. The table below shows actual gain using this transformer with some typical equipment output and input impedances (Z is impedance).

Consumer Output Z ▼	Pro Balanced Input Z		
	10 kÙ (625 Ù)	20 kÙ (1.25 kÙ)	40 kÙ (2.5 kÙ)
200 Ù	8.6 dB	9.7 dB	10.3 dB
500 Ù	5.9 dB	8.1 dB	9.4 dB
1 kÙ	**2.7 dB**	**5.9 dB**	**8.1 dB**

Not only will gain usually be much less than 12 dB, the load reflected to the consumer output (shown in parentheses) may cause headroom loss, increased distortion, and poor low-

Figure 41 - Balanced Output to Unbalanced Input
Top = None, Center = Output, Bottom = Input

frequency response. Often the only semi-technical description of a consumer output is "10 kÙ minimum load." It is futile to increase the turns ratio of the transformer in an attempt to overcome the gain problem — it only makes the reflected loading losses worse! In most situations, a 1:1 transformer can be used because the "pro" equipment can easily provide the required gain. Of course, a 1:1 *input* transformer will provide far superior noise immunity from ground loops.

The point here is that the noise rejection provided by an input transformer with a Faraday-shield is far superior to that provided by an output type. But the input transformer must be used at the receiver or destination end of an interface cable. In general, input transformers can drive no more than two feet of typical shielded cable — the capacitance of longer cables will erode their high-frequency bandwidth. Although output type (no Faraday shield) transformers are not as good at reducing noise, their advantage is that they can be placed anywhere along an interface cable, at the driver end, at a patch-bay, or at the destination end, and work equally well. In all the test cases discussed in this section, results of using *both* an output and an input transformer produced results *identical* to those using only an input transformer. For example, an unbalanced output does not need to be "balanced" by a transformer before transmission through a cable (this is a corollary of the balance versus symmetry myth), it needs only an input transformer at the receiver. There is rarely a need to use both types on the same line.

Definitions:

"Balanced Output" means a normal, non-floating source having a differential output impedance of 600 Ù and common-mode output impedances of 300 Ù, matched to within ± 0.1%;

"Balanced Input" means a typical electronically-balanced stage (an "instrumentation" circuit using 3 op-amps) having a differential input impedance of 40 kÙ and common-mode input impedances of 20 kÙ, trimmed for a CMRR over 90 dB when directly driven by the above "Balanced Output";

"Unbalanced Output" means a ground-referenced output having an output impedance of 600 Ù. This is representative of typical consumer equipment;

"Unbalanced Input" means a ground-referenced input having an input impedance of 50 kÙ. This is representative of typical consumer equipment;

"No Transformer" means a direct wired connection;

"Output Transformer" means a Jensen JT-11-EMCF (a popular line output transformer); and

"Input Transformer" means a Jensen JT-11P-1 (our most popular line input transformer).
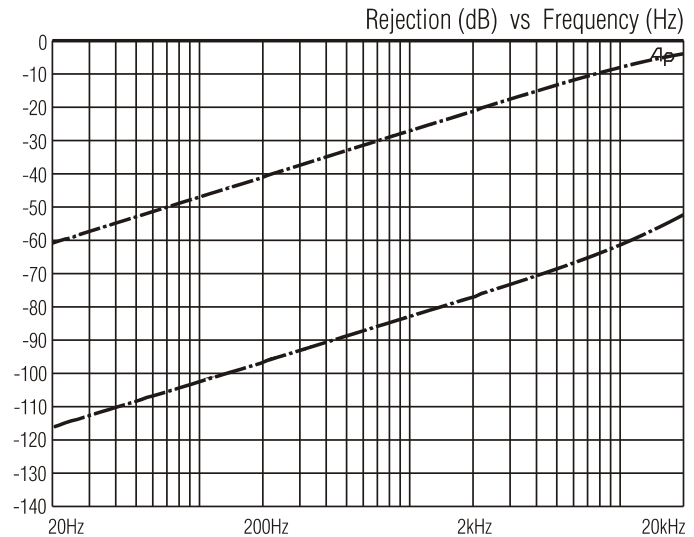
Audio Transformers

Handbook for Sound Engineers, 3rd Edition

2.2.5   *Speaker Distribution or "Constant Voltage"*

When a number of low-impedance speakers are located far from a power amplifier, there are no good methods to interconnect them in a way that properly loads the amplifier. The problem is compounded by the fact that power losses due to the resistance of the inter-connecting wiring can be substantial. The wire gauge required is largely determined by the current it must carry and its length. Borrowing a technique from power utility companies, boosting the distribution voltage reduces the current for a given amount of power and allows smaller wire to be used in the distribution system. *Step-down* "matching" transformers, most often having taps to select power level and/or speaker impedance, are used at each location. This scheme not only reduces the cost of wiring but allows system designers the freedom to choose how power is allocated among the speakers. These so-called "constant-voltage" speaker distribution systems are widely used in public address, paging, and background music systems. Although the most popular is 70-volt, others include 25-volt, 100-volt, and 140-volt. Because the higher voltage systems offer the lowest distribution losses for a given wire size, they are more common in very large systems. It should also be noted that only the 25-volt system is considered "low-voltage" by most regulatory agencies and the wiring in higher voltage systems may need to conform to *ac power* wiring practices.

It is important to understand that these nominal voltages exist on the distribution line only when the driving amplifier is operating at full rated power. Many specialty power amplifiers have outputs rated to drive these lines directly but ordinary power amplifiers rated to drive speakers can also drive such lines, according to the following table:

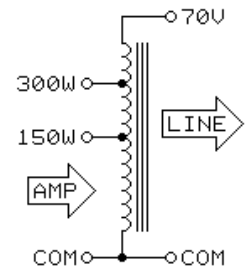| Amplifier Rated Output, Watts			**Output**
at 8 Ù	at 4 Ù	at 2 Ù	**Voltage**
1,250	2,500	5,000	**100**
625	1,250	2,500	**70.7**
312	625	1,250	**50**
156	312	625	**35.3**
78	156	312	**25**

For example, an amplifier rated to deliver 1,250 watts of continuous average power into an 8 Ù load could drive a 70-volt distribution line *directly* as long as the sum of the power delivered to all the speakers doesn't exceed 1,250 watts. Although widely used, the term *"rms watts"* is technically ambiguous [5]. In many cases, the benefits of constant-voltage distribution are desired, but the total

Figure 42 - Step-Up Auto-Transformer

power required is much less. In that case a *step-up* transformer can be used to increase the output voltage of an amplifier with less output. This is often called "matching" it to the line because such a transformer is actually transforming the equivalent line impedance down to the rated load impedance for the amplifier. Most of these step-up transformers will have a low turns ratio. For example, a 1:1.4 turns ratio would increase the 50-volt output to 70 volts for an amplifier rated at 300 watts into 8 Ù. In such low-ratio applications, the *auto-transformer* discussed in section 11.1.2.2 has cost and size advantages. **Figure 42** is a schematic of an auto-transformer with taps for turns ratios of 1:1.4 or 1:2 which could be used to drive a 70-volt line from amplifiers rated for either 300 or 150 watts respectively at 8 Ù. Several power amplifier manufacturers offer such transformers as options or accessories.

A "line to voice-coil" transformer is usually necessary to step-down the line voltage and produce the desired speaker power:

Speaker Power in Watts			**Speaker**	Transformer Step-Down Turns Ratio Required			
16 Ù	8 Ù	4 Ù	**Volts**	100 V	70 V	35 V	25 V
32	64	128	**22.63**	4.42	3.12	1.56	1.10
16	32	64	**16**	6.25	4.42	2.21	1.56
8	16	32	**11.31**	8.84	6.25	3.12	2.21
4	8	16	**8**	12.5	8.84	4.42	3.12
2	4	8	**5.66**	17.7	12.5	6.25	4.42
1	2	4	**4**	25	17.7	8.84	6.25
0.5	1	2	**2.83**	35.3	25	12.5	8.84
0.25	0.5	1	**2**	50	35.3	17.7	12.5
0.125	0.25	0.5	**1.41**	71	50	25	17.7

These step-down transformers can be designed several ways.
Figure 43 shows a design where the line voltage is selected at the
primary side and the power level is selected at the secondary while
Figure 44 shows a design where power level is selected on the
primary side and speaker impedance is selected at the secondary.
As clear from the repeating patterns in the table above, there are
many combinations of line voltage, speaker impedance, and power
level that result in the same required turns ratio in the matching
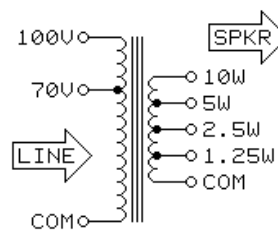transformer.

Figure 43 - Transformer
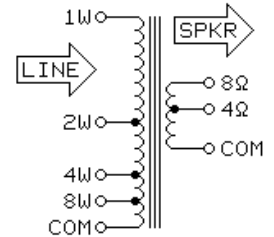with Secondary Taps for
Power Selection

Figure 44 - Transformer
with Primary Taps for
Power Selection

Since the constant-voltage line has a very low source impedance,
and the transformer is loaded by a low-impedance loudspeaker,
transformer high-frequency response is usually not a design issue.
As in any transformer, low-frequency response is determined by primary inductance and total source impedance, which is
dominated by the primary winding resistance since the driving source impedance is very low. Winding resistances of both primary
and secondary contribute to insertion loss. In efforts to reduce size and cost, the fewest turns of the smallest wire possible are often
used, which raises insertion loss and degrades low-frequency response. Generally, an insertion loss of 1 dB or less is considered
good and 2 dB is marginally acceptable for these applications.

It is very important to understand that, while the frequency response of a transformer may be rated as ‑1 dB at 40 Hz, its rated
power does NOT apply at that frequency. Rated power, or maximum signal level is discussed in section 1.3.1. In general, level
handling is increased by more primary turns and more core material and it takes more of both to handle more power at lower
frequencies. This ultimately results in physically larger, heavier, and more expensive transformers. When any transformer is driven
at its rated level at a lower frequency than its design will support, core saturation is the result. The sudden drop in permeability of
the core effectively reduces primary inductance to zero. The transformer primary now appears to have only the dc resistance of its
winding, which may be only a few ohms. In the best scenario, some ugly-sounding distortion will occur and the line amplifier will
simply current limit. In the worst scenario, the amplifier will not survive the inductive energy or "kick" fed back to it as the
transformer comes out of saturation. This can be especially dangerous if large numbers of transformers saturate simultaneously.

In 1953, the power ratings of "speaker matching transformers" were based on 2% distortion at 100 Hz [6]. Traditionally, the
normal application of these transformers has been speech systems and this power rating standard assumes very little energy will
exist under 100 Hz. The same reference recommends that transformers used in systems with "emphasized bass" should have
ratings higher than this 100-Hz "nominal power" rating and those used "to handle organ music" should have ratings of at least four
times nominal. Since the power ratings for these transformers is rarely qualified by a specification stating the applicable frequency,
it seems safe to assume that ***the historical 100 Hz power rating applies to most commercial transformers available today***.

If a background music system, for example, requires good bass response, it is wise to use over-rated transformers. Reducing the
voltage on the primary side of the transformer will extends its low-frequency power handling. Its possible, using the table above,
to use different taps to achieve the same ratio while driving less than nominal voltage into the transformer primary. For example, a
70-volt line could be connected to the 100-volt input of the transformer in **Figure 33** and, for example, the 10-watt secondary tap
used to actually deliver 5 watts. In any constant-voltage system, saturation problems can be reduced by appropriate high-pass
filtering. Simply attenuate low-frequency signals before they can reach the transformers. In voice-only systems, problems that arise
from breath pops, dropped microphones, or signal switching transients can be effectively eliminated by a 100-Hz high-pass filter
ahead of the power amplifier. In music systems, attenuating frequencies too low for the speakers to reproduce can be similarly
helpful.

2.2.6   *Telephone Isolation or "Repeat Coil"*

In telephone systems it was sometimes necessary to
isolate a circuit which was grounded at both ends. This
"metallic circuit" problem was corrected with a "repeat
coil" to improve "longitudinal balance." Translating
from telephone lingo, this balanced line had poor
common-mode noise rejection which was corrected with
a 1:1 audio isolation transformer. The Western Electric
111C repeat coil was widely used by radio networks and
others for high-quality audio transmission over 600 Ù
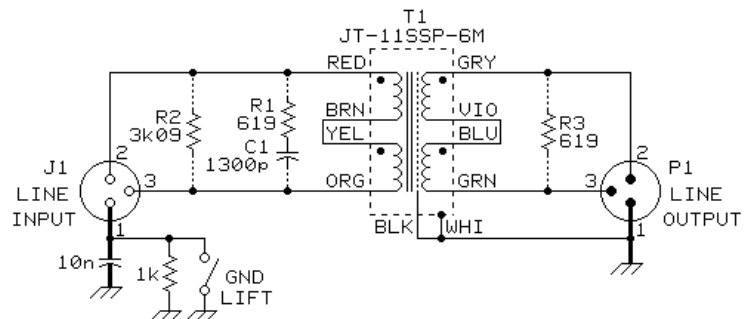phone lines. It has split primary and secondary windings

Figure 45 - "Repeat Coil" Ground Isolation for 600 Ù Lines

and a Faraday shield. Its frequency response was 30 Hz to 15 kHz and it had less than 0.5 dB insertion loss. Split windings allow them to be parallel connected for 150 Ù use.

Figure 45 shows a modern version of this transformer as a general purpose isolator for low-impedance circuits, such as in a recording studio patch-bay. Optional components can be useful in some applications. For example, network R1 and C1 will flatten the input impedance over frequency, R2 will trim the input impedance to exactly 600 Ù, and R3 can be used to properly load the transformer when the external load is high-impedance or "bridging."

2.2.7   *Telephone Directional Coupling or "Hybrid"*

Telephone "hybrid" circuits use bridge nulling principles to separate signals which may be transmitted and received simultaneously on a 2-wire line. This nulling depends critically on well-controlled impedances in all branches of the circuits. This nulling is what suppresses the transmit signal (your own voice) in the receiver of your phone while allowing you to hear the receive signal (the other party).

A two-transformer hybrid network is shown in **Figure 46**. The arrows and dashed lines show the current flow for a signal from the transmitter TX. Remember that the dots on the transformers show points having the same instantaneous polarity. The transformer turns ratios are assumed to be 1:1:1. When "balancing network" Z_N has an impedance that matches the line impedance Z_L at all significant frequencies, the currents in the Z_L loop (upper) and Z_N loop (lower) will be equal. Since they flow in opposite directions in the RX transformer (right), there is cancellation and the TX signal does not appear at RX. A signal originating from the line rather than TX is not suppressed and is heard in RX. A common problem with hybrids of any kind is adjusting network Z_N to match the telephone line, which may vary considerably in impedance even over relatively short time spans.
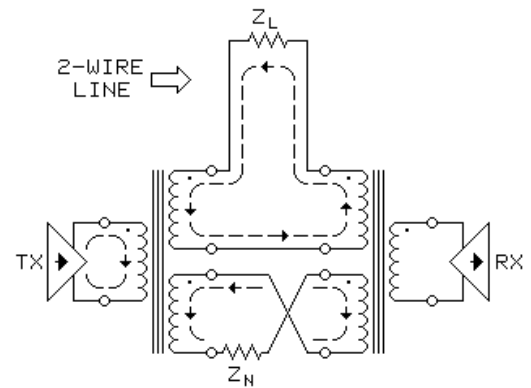

Figure 46 - Two-Transformer "Hybrid"

If the transmitter and receiver are electrically connected, the single transformer method, shown in **Figure 47**, can be used. Any well-designed transformers with accurate turns ratio can be used in hybrid applications.
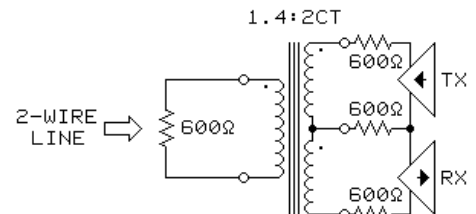
2.2.8   *Moving-Coil Phono Step-Up*

Outboard boxes are sometimes used to adapt the output of low-output, low-impedance moving-coil phono pickups to pre-amplifier inputs intended for the more ordinary high-impedance moving-magnet pickups. These pre-amplifiers have a standard input impedance of 47 kÙ. **Figure 48** shows a 1:37 step-up transformer used for this purpose. It has a voltage gain of 31 dB and reflects its 47 kÙ pre-amplifier load to the pickup as about 35 Ù. This keeps loading loss on the pickup to about 1 dB. The series RC network on the secondary provides proper damping for smooth frequency response. Double magnetic shield cans are used because of the very low signal levels involved and the low-frequency gain inherent in the RIAA playback equalization. In these applications, it is extremely important to keep all leads to the pickup tightly twisted to avoid hum from ambient magnetic fields.
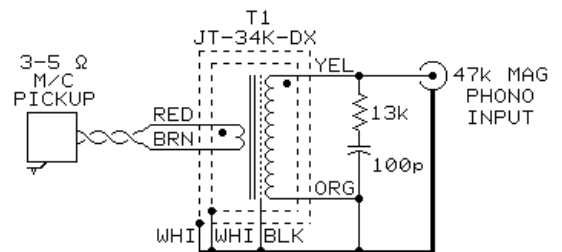

Figure 47 - Single-Transformer "Hybrid"

Figure 48 - Step-Up Transformer for
Moving Coil Phono Pickup

3   Measurements and Data Sheets

3.1 Testing and Measurements

3.1.1   *Transmission Characteristics*

The test circuits below are the basic setups to determine the signal transmission characteristics of output and input type transformers, respectively, shown in the diagrams as DUT for "device under test." In each case, the driving source impedance must be specified and is split into two equal parts for transformers specified for use in balanced systems. For example, if a 600 Ω balanced source is specified, the resistors Rs/2 become 300 Ω each. The generator indicated in both diagrams is understood to have symmetrical voltage outputs. The buffer amplifiers shown are used to provide a zero source impedance, which is not available from most commercial signal sources. The generator could be used in an unbalanced mode by simply connecting the lower end of the DUT primary to ground. The specified load impedance must also be placed on the secondary. For output transformers, the load and meter are often floating as shown in **Figure 49**. For input transformers, a specified end the secondary is generally grounded as shown in **Figure 50**.
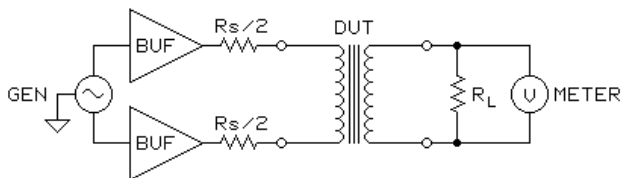
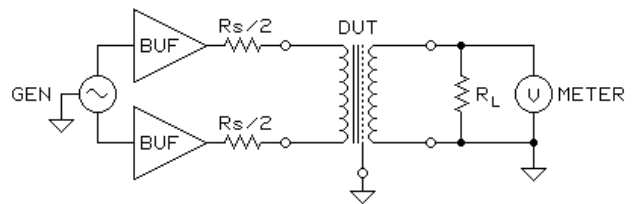**Figure 49** - Transmission Tests for Output Types

Figure 50 - Transmission Tests for Input Types

These test circuits can be used to determine voltage gain or loss (turns ratio when R_L is infinite), frequency response, and phase response. If the meter is replaced with a distortion analyzer, distortion and maximum operating level may be characterized. Multi-purpose equipment such as the Audio Precision System 1 or System 2 can make such tests convenient. Testing of high-power transformers usually requires an external power amplifier to boost the generator output as well as some hefty power resistors to serve as loads.

3.1.2   *Balance Characteristics*

Tests for common-mode rejection are intended to apply a common-mode voltage through some specified resistances to the transformer under test. Any differential voltage developed then represents undesired conversion of common-mode voltage to differential mode by the transformer. In general terms, CMRR or common-mode rejection ratio, is the ratio of the response of a circuit to a voltage applied normally (differentially) to that same voltage applied in common-mode through specified impedances. This conversion is generally the result of mismatched internal capacitances in the balanced winding. For output transformers, the most common test arrangement is shown in **Figure 51**. Common values are 300 Ω for RG and values from zero to 300 Ω for Rs/2. Resistor pairs must be very well matched.
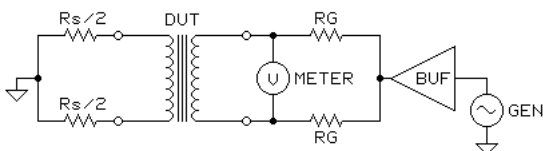
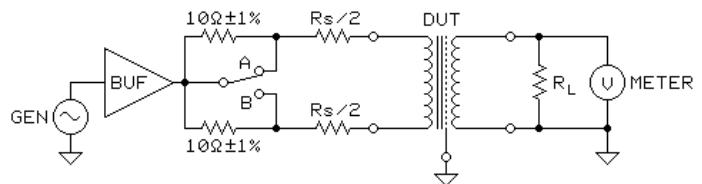**Figure 51** - Common-Mode Test for Output Types

Figure 52 - IEC Common-Mode Test for Input Types

Traditionally, CMRR tests of balanced input stages involved applying the common-mode voltage through a pair of very tightly-matched resistors. As a result, such traditional tests were not accurate predictors of real-world noise rejection in some very widely used electronically-balanced inputs. The IEC recognized this a number of years ago and solicited help to revise the test. The problem arises from the fact that the common-mode output impedances of "balanced" sources in typical commercial equipment are <u>not</u> matched with laboratory precision. Imbalances of 10 Ω are quite common. This author, through an educational process about balanced interfaces in general, suggested a more realistic test which was adopted by the IEC in their document 60268-3 "Testing of Amplifiers" in August, 2000. The "Informative Annex" of this document is a concise short-course explaining the nature of a balanced interface. The method of the new test, as shown in Figure **52**, is simply to introduce a 10 Ω imbalance, first in one line and then in the other. The CMRR is then computed based on the highest differential reading observed.

3.1.3 Resistances, Capacitances, and Other Data

Other data which can be very helpful to an equipment or system designer includes resistances of each winding and capacitances from winding to winding or winding to Faraday shield or transformer frame. Do not use an ohmmeter to check winding resistances unless you are able to later demagnetize the part. Ordinary ohmmeters, especially on low-ohm ranges, can weakly magnetize the core. If an ohmmeter simply must be used, use the highest ohm range (where the test current is least).
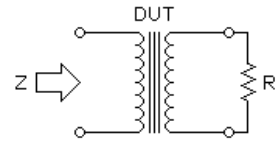
Figure 53 - Impedance Tests

Capacitances are usually measured on impedance bridges and, to eliminate the effects of winding inductances, with all windings shorted. Total capacitances can be measured this way, but balance of capacitances across a winding must be measured indirectly. CMRR tests are effectively measuring capacitance imbalances.

As shown in **Figure 53**, sometimes the input impedance of a winding is measured with specified load on other windings. This test includes the effects of primary resistance, secondary resistance, and the parallel loss resistance RC shown in **Figure 8** and **Figure 13**. If specified over a wide frequency range, it also includes the effects of primary inductance and winding capacitances.

Breakdown voltages are sometimes listed as measures of insulation integrity. This is normally done with special equipment, sometimes called a "hi-pot" tester, which applies a high voltage while limiting current to a very low value.

3.2 Data Sheets

3.2.1 Data to Impress or to Inform?

As with other products, many data sheets and other product specifications are designed to impress rather than inform. Specifications offered with unstated measurement conditions are essentially meaningless, so **a degree of skepticism is always appropriate** before comparisons are made. A few examples:

"Hum Eliminator" and "Line Level Shifter" products with no noise rejection specs of any kind!

"Line Level Shifter" products with no gain spec at all! Section 2.2.4 explained why you likely never see one.

"Maximum Power" or "Maximum Level" listed with <u>no</u> frequency and <u>no</u> source impedance specified!

Other specifications, while true, may mislead those not wise in the ways of transformers.

"Maximum Level" and "Distortion" are commonly specified at 50 Hz, 40 Hz, 30 Hz, or 20 Hz. Be careful, the 50 Hz specs will always be much more impressive than those at 20 Hz! There is an approximate 6 dB per octave relationship at work here. A transformer specified for level or distortion at 40 Hz for example, will handle about 6 dB *less* level at 20 Hz and have *at least* twice the distortion!

Seen in advertising copy: "Frequency response 10 Hz to 40 kHz ±1 dB into 10 kÙ load" and "Distortion less than 0.002% at 1 kHz." What about the source impedance? Response at 10 Hz and low distortion is a lot easier from a 0 Ù source than from a 1 kÙ source — and 1 kHz is not a very revealing frequency for distortion tests. Section 1.3.1 explains.
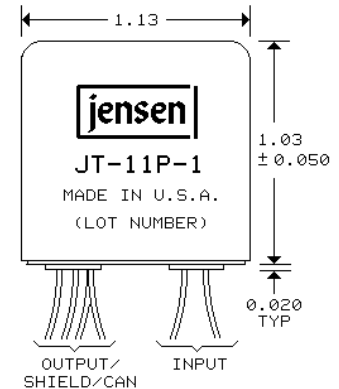
3.2.2 Comprehensive Data Sheet Example

For reference, the following is offered as a sample of a data sheet that has been called truly useful and brutally honest.
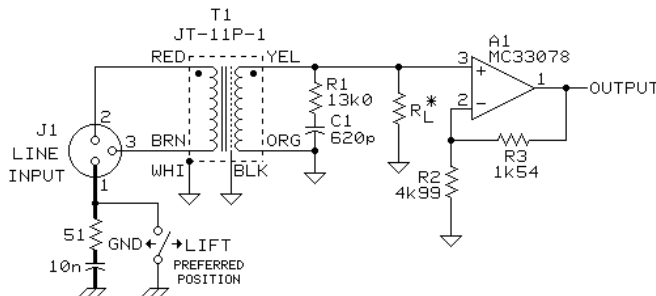
JT-11P-1

LINE INPUT TRANSFORMER
1:1 FOR "BALANCED BRIDGING" INPUTS

● *Ideal for balancing any high-impedance unbalanced input*
● *Wide bandwidth: −3 dB at 0.25 Hz and 100 kHz*
● *Recommended for levels up to +20 dBu at 20 Hz*
● *High input impedance: 13 kΩ with 10 kΩ load*
● *High common-mode rejection: 107 dB at 60 Hz*

This transformer is designed for use in wideband line input stages. Distortion remains very low and CMRR remains high, even when driven by high source impedances. The primary is fully balanced and its leads may be reversed to invert polarity, if required. A 30 dB magnetic shield package is standard.

```
1.13
jensen
JT-11P-1
MADE IN U.S.A.
(LOT NUMBER)
1.03
±0.050
0.020
TYP
OUTPUT/        INPUT
SHIELD/CAN
```

#30 AWG (7x38) UL STYLE 1061 COLOR CODED WIRE LEADS, 8" MINIMUM LENGTH

```
T1
JT-11P-1          A1
RED------YEL    3  MC33078
                +        1
        R1        *      ──OUTPUT
J1      13k0  RL  2  −
LINE    C1
INPUT  3 BRN  ORG 620p      R3
                            1k54
    2
    1 WHI  BLK      R2
                    4k99
51  GND←/→LIFT
10n    PREFERRED
       POSITION
```

*RL MUST BE 10 kΩ OR GREATER
OMIT DAMPING NETWORK R1 AND C1 FOR RL = 10 kΩ ONLY

LOW NOISE UNITY GAIN INPUT STAGE
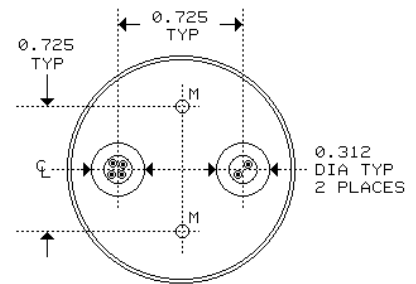
```
0.725
TYP
0.725
TYP
        M
CL              0.312
                DIA TYP
                2 PLACES
        M
BOTTOM VIEW
```
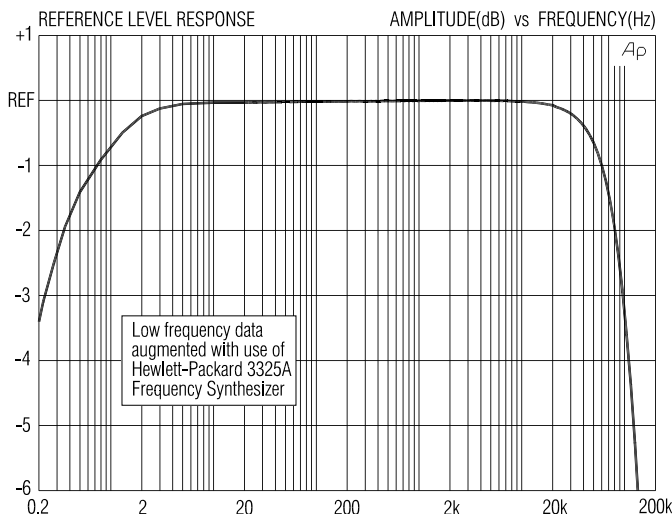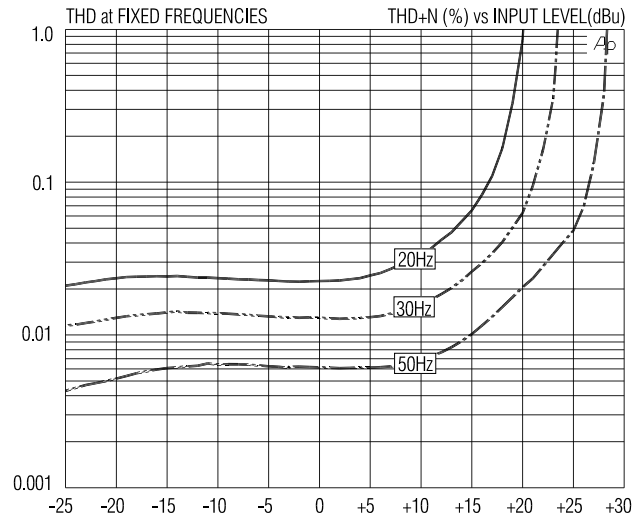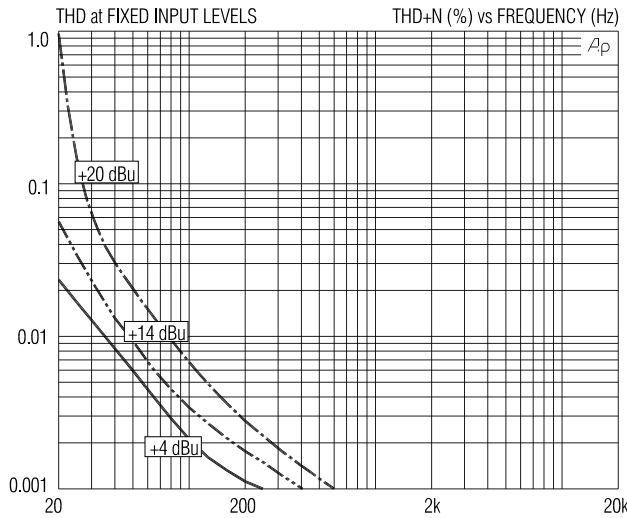
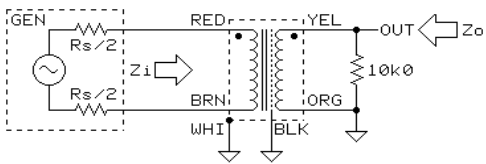USE ONLY #4 TYPE B SELF TAPPING SCREWS IN HOLES "M". ALLOW NO MORE THAN 0.15" PENETRATION INTO TRANSFORMER HOUSING.
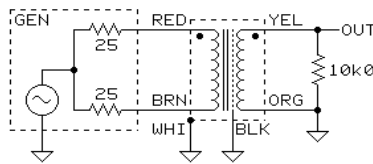
TYPICAL APPLICATION

REFERENCE LEVEL RESPONSE   AMPLITUDE(dB) vs FREQUENCY(Hz)

Low frequency data augmented with use of Hewlett-Packard 3325A Frequency Synthesizer

DEVIATION FROM LINEAR PHASE   DEVIATION (Deg) vs FREQUENCY (Hz)

THD at FIXED INPUT LEVELS   THD+N (%) vs FREQUENCY (Hz)

THD at FIXED FREQUENCIES   THD+N (%) vs INPUT LEVEL(dBu)
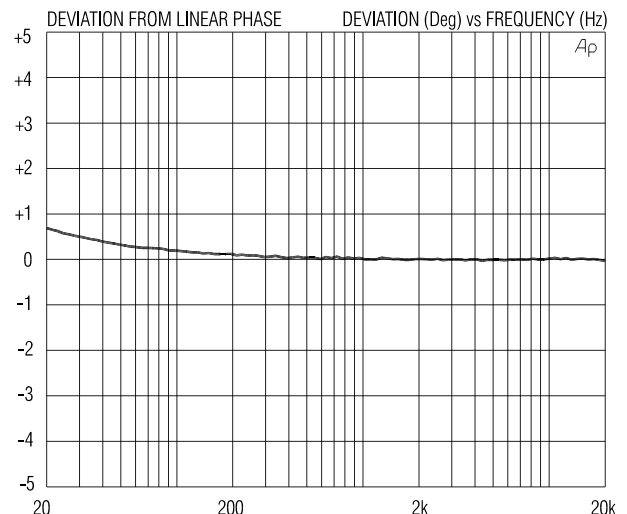
JT-11P-1 SPECIFICATIONS (all levels are input unless noted)

PARAMETER	CONDITIONS	MINIMUM	TYPICAL	MAXIMUM
Input impedance, Zi	1 kHz, +4 dBu, test circuit 1	12.3 kΩ	13.0 kΩ	13.7 kΩ
Voltage gain	1 kHz, +4 dBu, test circuit 1	−2.6 dB	−2.3 dB	−2.0 dB
Magnitude response, ref 1 kHz	20 Hz, +4 dBu, test circuit 1, Rs=600 Ω	−0.15 dB	−0.04 dB	0.0 dB
	20 kHz, +4 dBu, test circuit 1, Rs=600 Ω	−0.15 dB	−0.05 dB	0.0 dB
Deviation from linear phase (DLP)	20 Hz to 20 kHz, +4 dBu, test circuit 1, Rs=600 Ω		+0.6°	±2.0°
Distortion (THD)	1 kHz, +4 dBu, test circuit 1, Rs=600 Ω		<0.001%	
	20 Hz, +4 dBu, test circuit 1, Rs=600 Ω		0.025%	0.10%
Maximum 20 Hz input level	1% THD, test circuit 1, Rs=600 Ω	+18 dBu	+20 dBu	
Common-mode rejection ratio (CMRR) 50 Ω balanced source	60 Hz, test circuit 2		107 dB	
	3 kHz, test circuit 2	65 dB	73 dB	
Common-mode rejection ratio (CMRR) 600 Ω unbalanced source	60 Hz, test circuit 3		100 dB	
	3 kHz, test circuit 3		68 dB	
Output impedance, Zo	1 kHz, test circuit 1, Rs=50 Ω		2.34 kΩ	
DC resistances	primary (RED to BRN)		1.45 kΩ	
	secondary (YEL to ORG)		1.55 kΩ	
Capacitances @ 1 kHz	primary to shield and case		98 pF	
	secondary to shield and case		110 pF	
Turns ratio		0.999:1	1.000:1	1.001:1
Temperature range	operation or storage	0° C		70° C
Breakdown voltage (see IMPORTANT NOTE below)	primary or secondary to shield and case, 60 Hz, 1 minute test duration	250 V RMS		

TEST CIRCUIT 1   TEST CIRCUIT 2   TEST CIRCUIT 3

1/01

4   Installation and Maintenance

4.1 A Few Installation Tips

- Remember that there are very tiny wires inside an audio transformer. Its wire leads should never be used like a handle to pick it up. The internal bonds are strong, but one strong tug might result in an open winding.

- Be careful with sharp tools. A gouge through the outer wrapper of an output transformer can nick or cut an internal winding.

- Use either the supplied screws or ones no longer than recommended to mount transformers in shield cans. If the screws are too long, they'll bore right into the windings — big problem!

- Be careful about using magnetized tools. If a screwdriver will pick up a paper clip, it shouldn't be used to install an audio transformer.

- Don't drop a transformer. It can distort the fit of the laminations in output transformers and affect their low-frequency response. Mechanical stress (as in denting) of the magnetic shield can of an input transformer will reduce its effectiveness as a shield. For the same reason, don't over-tighten the clamp on transformers mounted with them.

- Twisting helps avoid hum pickup from ambient ac magnetic fields. This is especially true for mic level lines in splitters, for example. Separately twist the leads from each winding — twisting the leads from *all* windings together can reduce noise rejection or CMRR.

4.2 De-Magnetization

Some subtle problems are created when transformer cores and/or their shield cans become magnetized. Generally, cores become magnetized by having dc flow in a winding, even for a fraction of a second. It can leave the core weakly magnetized. Steel cores, because of their wider hysteresis loops, are generally the most prone to such magnetization. The only way to know if the core has some permanent magnetization is to perform distortion measurements. A transformer with an un-magnetized core will exhibit nearly pure third harmonic distortion, with virtually no even order harmonic distortion while magnetized ones will show significant even order distortion, possibly with 2nd harmonic even exceeding 3rd. A test signal at a level about 30 or 40 dB below rated maximum operating level at 20 or 30 Hz is typically the most revealing because it maximizes the contribution of hysteresis distortion.

Microphone input transformers used with phantom power are exposed to this possibility whenever a microphone is connected or disconnected from a powered input. However, distortion tests before and after exposure to the worst-case 7 mA current pulses have shown that the effects are indeed subtle. Third harmonic distortion, which normally dominates transformer distortions, is unaffected. Second harmonic, which normally is near the measurement threshold, is typically increased by about 20 dB but is still some 15 dB lower than the third harmonic. Is it audible? Some say yes. But even this distortion disappears into the noise floor above a few hundred Hz. In any case, it can be prevented by connecting and disconnecting microphones *only* when phantom power is off. However, such magnetized transformers can be de-magnetized.

Demagnetizing of low level transformers can generally be done with any audio generator having a continuously variable output It may take a booster of some sort to get enough level for output transformers (be sure there's no dc offset at its output!). The idea is to drive the transformer deeply into saturation (5% THD or more) and slowly bring the level down to zero. Saturation will, of course, be easiest at a very low frequency. How much level it takes will depend on the transformer. If you're lucky, the level required may not be hazardous to the surrounding electronics and the de-magnetizing can be accomplished without disconnecting the transformer. Start with the generator set to 20 Hz and its minimum output level, connect it to the transformer, then slowly (over a period of a few seconds) increase the level into saturation — maintain it for a few seconds — then slowly turn it back down to minimum. For the vast majority of transformers, this process will leave them in a demagnetized state.

Shield cans are usually magnetized by having a brief encounter with a strongly magnetized tool. Sometimes, transformers are unknowingly mounted on a magnetized chassis. When the shield can of an input transformer becomes magnetized, the result is *microphonic* behavior of the transformer. Even though quality input transformers are "potted" with a semi-rigid epoxy compound to prevent breakage of very fine wires, tiny movements between core and can activate what is essentially a variable reluctance microphone. In this case, a good strong tape head de-magnetizer can be used to de-magnetize the can. At the end of the production line, most transformers are routinely demagnetized with a very strong de-magnetizer just prior to shipment. Although I haven't tried it, I would expect that something like a degausser for 2" video tape (remember that!) would also de-magnetize even a large

steel-core output transformer.

References

[1]   Magnetic Shield Corporation, *Frequently Asked Questions*, www.magnetic-shield.com/faq.html.

[2]   Sowter, G.A.V., *Soft Magnetic Materials for Audio Transformers: History, Production, and Applications*, Journal of the Audio Engineering Society, October 1987, www.sowter.co.uk/pdf/GAVS.pdf.

[3]   Whitlock, Bill, *Balanced Lines in Audio: Fact, Fiction, and Transformers*, Journal of the Audio Engineering Society, June 1995, pp 454-464.

[4]   Smith, F. Langford, *Radiotron Designer's Handbook*, Wireless Press, Sydney, 4th Edition, 1953, p 208.

[5]   Woolf, Lawrence, *RMS Watt, or Not?*, Electronics World, December 1998, pp 1043-1045.

[6]   Smith, F. Langford, *op. cit.*, p 227.