Midcom Transformer Theory

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All science is either physics or stamp collecting.
—Ernest Rutherford (1871-1935)

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Introduction

This document describes Midcom products, many of which are inductive, most of which are transformers. While there are many kinds of transformers in the world, our focus will be limited to those products Midcom manufactures or those Midcom is capable of providing. While engineers are the intended audience here, anyone in the electronics industry may benefit from reading this document.

Of the many things our customers tell us, one thing we hear repeatedly from new design engineers is "I can't believe somebody hasn't designed a solid-state replacement for xxxx application." While it is true that some transformer applications have been replaced by silicon, there are many for which the transformer is the absolute best when it comes to price and performance. There are several reasons for this. One reason is that the transformer, being a passive "no-batteries-required" component, is convenient to use when no energy source is available to power a silicon device. Transformers of today are *not* the same as those of yesteryear—not any more than silicon product development stopped with the 555 timer. Modern transformers are smaller and cheaper, and perform better in ways that were only concepts a decade ago. As magnetic material research continues, the promise of sustained improvement moves right along with it. Switchmode power supplies less than 10 watts are now an economical replacement for their linear counterparts, certainly in terms of energy efficiency, but also product cost and particularly when small size is a requirement. Having pointed out that transformers are *not* "dead," I must also add that this document will be updated when we learn of new developments. If you have downloaded this copy from the Midcom web site, you may wish to check back with us to see if your copy is the most current version.

Another common statement made by our customers is, "Transformers are such a black art. I don't know how they work, so I'm wary of using them." I will attempt to dispel in the pages that follow most if not all of the mystique of the transformer. This ambitious goal of mine requires some exertion on your part as a willing reader of this document, but the reward will be worth your effort.

Lastly, I don't wish to present myself as an expert on magnetics, but only as an engineer who has spent most of his career thus far in the pursuit of knowledge in an area where there is much myth and few mentors.

Geometry of transformer construction

The transformer was invented by Michael Faraday in 1831, although it was called an *induction coil* at the time. People in the telephone industry still refer to induction coils, but the term transformer is universally understood today.

A transformer is defined as two inductors that happen to be magnetically coupled. This means that two conductors in reasonable proximity to one another will exhibit a "transformer" effect if one is carrying an appreciable amount of current.

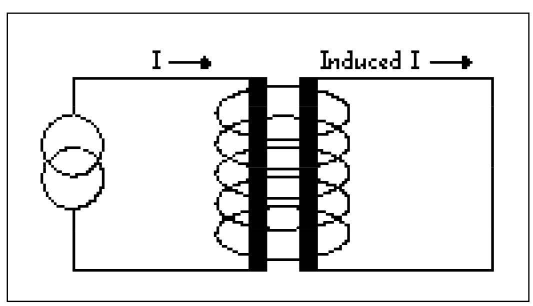


Figure 1

Two parallel conductors do not make a terribly efficient transformer. A better approach is to use a *coil*, usually created by attaching conductors to a *coil former* also known as a *bobbin* which is rotated to create the coil. A special class of wire called *magnetwire* has been developed to facilitate the construction of inductors and transformers. Magnetwire typically consists of copper wire with a thin coating of polyurethane insulation. The insulation coating must be evenly distributed around the perimeter of the wire. Conscientious wire manufacturers apply the insulation coating in several steps; this prevents uneven distribution of the insulation which in turn results in weak spots where the coating is too thin.

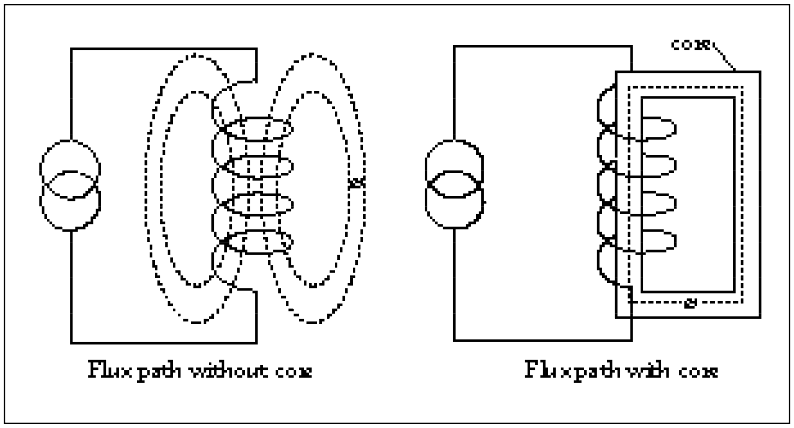


Figure 2

A coil of wire is able to concentrate a magnetic field by, not surprisingly, a multiplication factor equal to the of the number of turns of the coil. The variable N is usually reserved in magnetics equations to define the number of turns of the coil. Since N is essentially a multiplication factor, it is a dimensionless unit for the same reason percent and slope are dimensionless entities.

Two coils placed in relative proximity certainly have the ability to couple magnetic energy, but only to the extent that the energy leaving one coil can be captured by the other. If we wish to separate the coils by an appreciable distance, as we do when trying to provide an isolation barrier, we need something to contain the magnetic energy and route it through both coils with minimum leakage. We call this item a *core*. (See Figure 2.)

About magnetwire

Magnetwire in North America is defined by the American Wire Gauge system, or AWG. Unlike other systems, the AWG system is arranged in a progressive sequence where the bare wire diameter decreases by half every six wire gauge numbers. For example, AWG 36 bare wire is 0.005 inches in diameter, which means that AWG 30 is nominally 0.01 inches in diameter. This is convenient for those of us with poor memories, but it doesn't apply to insulated magnetwire—only bare wire.

The thickness of the insulation coating is related to the wire diameter, but not in a nice, linear relationship. Insulation thickness is defined by a NEMA (National Electrical Manufacturer's Association) specification, so you needn't worry that people are just making this up as they go along. The NEMA specification defines two insulation thicknesses for each gauge of wire: single (light) build and double (heavy) build. There are also IEC and JIS specifications for magnetwire which have three grades and four grades of insulation thicknesses, respectively.

A key attribute of magnetwire is that it is solderable, meaning that its insulation coating may be easily removed during a soldering process thus allowing an electrically sound connection between the copper wire and the terminal to which it is connected.

Of all the metals and alloys of metal that comprise magnetwire, copper is by far the most common and least expensive. Copper is also the second-best in the low resistance category. The only other element with lower resistivity is silver, but by only a few percent which hardly justifies its additional expense except in extremely critical applications.

Transformer and inductor cores may be made from various magnetic materials. The measure of a core material's effectiveness as a means of containing magnetic flux is known as *permeability*. The permeability of a material may be expressed in two ways: intrinsic and relative. *Relative permeability* is permeability expressed as a ratio of that over free space—a volume not occupied by anything remotely magnetic. You may have heard the term "air-core coil" which refers to a coil having no core other than free space or "air." (If a magnetics supplier tells you your air-core coils will be delivered late because the cores are on back order, they are pulling your leg. Either that, or you've got another, more serious problem on your hands.)

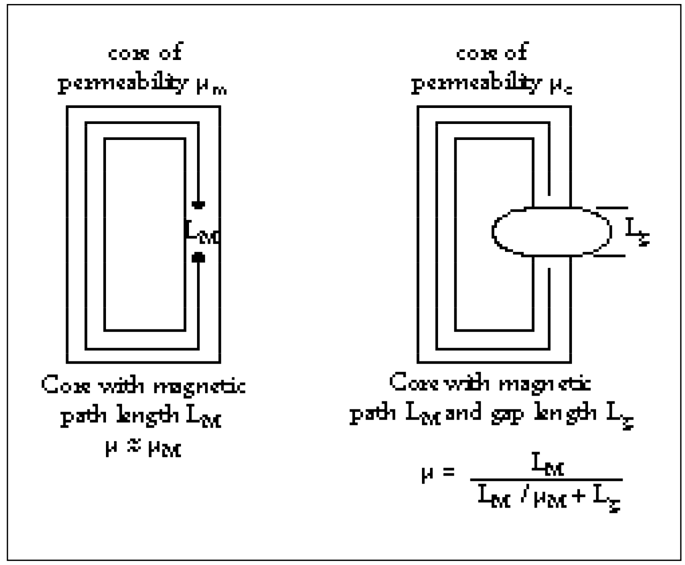


Figure 3

The permeability of free space, μ_0, has a value of $4\pi \times 10^{-7}$ henries/meter. This means that a thin conductor stretched into a straight line exactly one meter in length will have an inductance of about 1.26 microhenries. The permeability of a core material is expressed as the relative permeability, μ_R, of that material when compared to that of free space, thus $\mu_M = \mu_0 \mu_R$. Most of our materials have relative permeabilities in the range of a few hundred to 50,000 with the bulk ranging from 1000 to 10,000. *Intrinsic permeability* is the base permeability of a material in henries/meter.

We will cover various core shapes in a later chapter but for now we need only know that to make maximum use of a material's permeability, we must have a *closed path* for the magnetic flux to follow. In some cases we introduce a gap into the core's structure to reduce its *effective permeability* the same way that a high-value resistor may be used to limit current in an electrical circuit. We may choose to include an air gap in a core to prevent saturation with DC current, or to control the effects of temperature on inductance by reducing an inductor's dependence on material permeability. Figure 3 shows the effects of gapping a core structure.

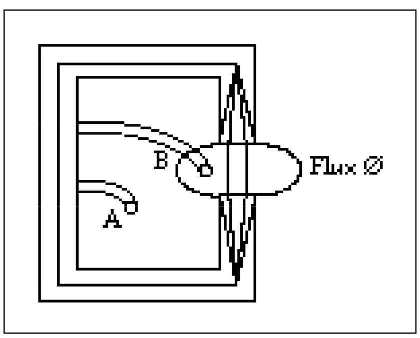


Figure 4

Figure 4 illustrates a magnetic phenomenon known as *fringing*. Fringing occurs when magnetic flux reaches a discontinuity—usually an air gap—in the core structure where the local permeability is far less than that of the material as a whole. When the flux reaches the gap, it "fringes" or bulges outward from it. This effect can be important if there are turns of wire near the gap. Each turn of wire must be completely encircled by magnetic flux if it is to be counted as an electrical turn. Flux will flow around complete turns near the gap, rendering them ineffective in a magnetic sense. Wire A is completely encircled by flux Δ whereas wire B is only partially encircled. It is important to remember that flux is actually a gradient and not made up of lines which are shown for illustrative purposes only.

Polarity

Transformers do an excellent job of isolating two parts of a circuit. In fact, a transformer is the least expensive way of providing *galvanic isolation* between two circuits. Although the input winding of a properly-constructed transformer may have a "hot" side and a "ground" side, the output winding may "float," having no reference to the input side ground. In other words, if you were to connect a voltmeter between either output terminals to the ground connection on the input side, you would read little or no voltage. This comes as no surprise to anyone familiar with transformer operation, but most circuit analysis programs have no convenient way of dealing with a floating voltage. You may find that your favorite analysis package will refuse to generate results until you reference one of the output windings to ground.

Although the input and output windings may have no electrical connection, there is still a relationship between them. The transformer industry has informally decided to refer to the input side of a transformer as the *primary* and the output side as the *secondary*. It is possible and indeed sometimes desirable to have multiple primary and secondary windings on the same transformer.

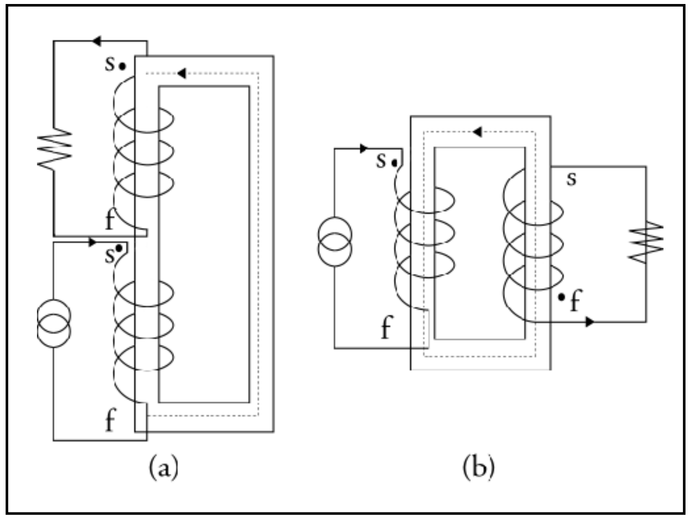


Figure 5

We define the *polarity* of a transformer winding through the use of a *dot convention* where current flowing into the dotted side of a transformer will produce a current flow out of the dotted side at its output. This is illustrated in Figure 5.

The method of winding a coil on a rotating arbor dictates that the coil have a *start* terminal and a *finish* terminal. The start and finish terminals of each winding are denoted with the letters "s" and "f" respectively in Figure 5. In Figure 5(a) we see that the magnetic flux flows into the start terminals of each winding, thus causing each start to have the same polarity and allowing us to place the polarity dots at each winding's start termination. In Figure 5(b) flux again flows into the start of the input winding, but it flows into the *finish* of the output winding. This reversing of the flux causes current to flow out of the finish termination—the opposite of the case shown in (a). This illustrates an important point about coil winding: start and finish terminations of different windings will have the same polarity only if they have the same orientation with respect to magnetic flux. To aid in understanding this principle, think of the second winding being slid around the core structure such that it is superimposed upon the first. If the start of the second winding follows the same path around the core as the first winding, the start terminations will have the same polarity (and by default, so will the finish terminations). The converse is also true: if the first winding's start termination direction is opposite that of the second, the start of the first winding will have the same polarity as the *finish* of the second.

This can be confusing, but thankfully most coils are wound such that reversals are rarely needed. Thus in most cases, each of the winding starts of a given transformer will have the same polarity sense. Notable exceptions are common-mode chokes wound on toroidal cores where winding mirror symmetry improves common-mode rejection, and double-coil transformers using U-U and U-I cores where one coil may be easily reversed with respect to the other. Winding reversal is also employed when longitudinal balance must be equal when either the start or the finish of the secondary may be referenced to ground.

Many transformer manufacturers have adopted the practice of denoting the winding starts and finishes wth a polarity dot. But when coils are arranged such that the winding sense between two coils is exactly opposite, we must apply one of the dots to a finish winding. This is required to assure that the polarity dots convey the correct phasing information. This is covered later, in the section on U- and C-shaped cores in Chapter 4, "Construction."

Chapter 1 Inductor Modeling

Before we begin to work with the entire transformer equivalent circuit, we should understand something about inductors. Pure inductances are fictional, just as are pure resistances and pure capacitances. Each has elements of the other two, inextricably connected in a mixture we can only hope to model over a confined frequency range. Outside the range of frequencies, all bets are off and a new model must be chosen. Remember that a capacitance appears between any two points in a circuit where a voltage differential exists; an inductance appears between any two points in a circuit where a current flows; and a resistance appears when a current flowing in a circuit where a current flows and a voltage differential exists. Those statements are just profound enough to warrant bullet points:

- A capacitance appears between any two points in a circuit where a voltage differential exists.
- An inductance appears between any two points in a circuit where a current flows.
- A resistance appears between any two points in a circuit where a current flows and a voltage differential exists.

On the surface, these seem to be obvious statements of the laws of circuits. Because we don't have perfect insulators and conductors, even a very simple circuit will be fraught with parasitic elements of resistance, capacitance and inductance. For example, a very complete, but hard-to-analyze equivalent circuit for an inductor is shown here:

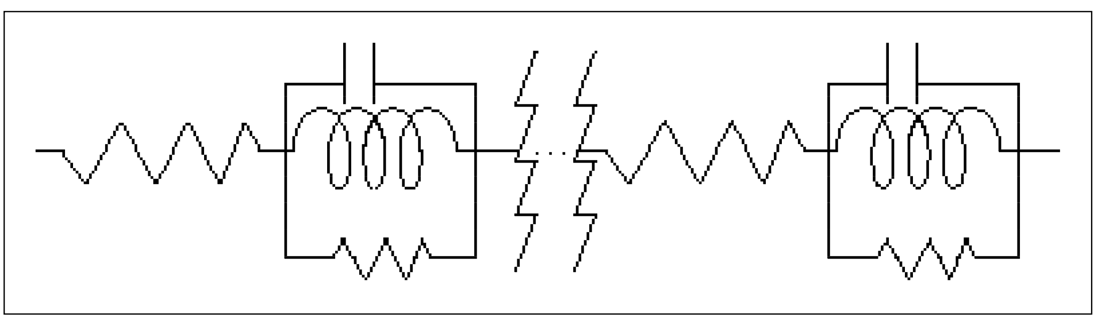


Figure 6

Building up a suitable model, a little at a time

Figure 7 shows a first-order approximation of an inductor.

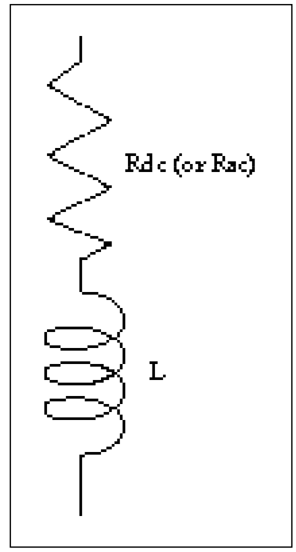


Figure 7

We have separated the inductance and resistance and placed them in series. This fits our view of an inductor consisting of a coil of wire having a measurable resistance. Since an inductor's reactance is proportional to the frequency of the applied excitation signal, we can measure the resistive portion separately by applying a zero-hertz, or DC signal and measuring the resulting voltage drop across the entire device. In the magnetics trade, we call this series resistance the *DC Resistance*, or *DCR*. Measuring DCR is an easy and quick way to assure that the coil is wound with the correct wire gauge and turns count, and that it has no obvious defects, such as a large number of shorted turns. Measuring DCR is *not* a reliable way of determining whether a coil has a small number of shorted turns. Means exist to detect small numbers of shorted turns and will be covered in a later chapter. DC resistance is sometimes denoted by *Rdc* in keeping with standard electronics industry nomenclature.

Element change to due frequency and flux density

The coil resistance can change as a function of the frequency due to the *skin effect*. The skin effect is named after the effect by which flux linkages force the current distribution in a conductor's cross section to occupy the region nearest the outer surface of the conductor. The reduction in effective cross-sectional area causes an increase in winding resistance. To differentiate between coil resistance at DC and coil resistance at a frequency where skin effect becomes noticeable, we denote the coil resistance as Rac when skin effect is present and accounted for. The *skin depth* defines the effective conductor depth and describes the dimensions of the effective current-carrying area. Skin depth is strictly a function of frequency, which is why R_{COIL} is shown to be dependent solely on frequency, f, and not upon flux density, B.

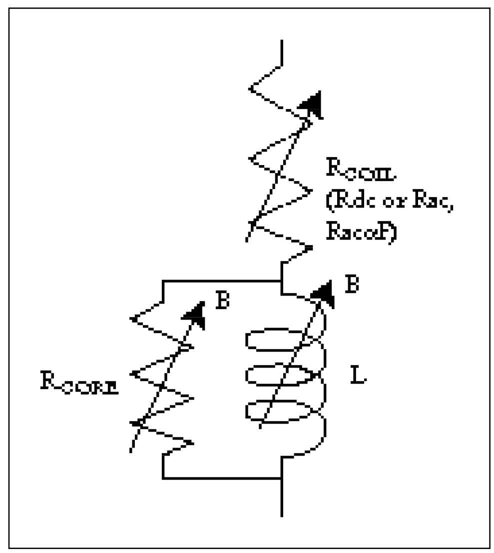


Figure 8

Losses in the inductor's core result in a parasitic resistance which may be modeled as a resistor in parallel with the inductor. The parallel resistance and inductance are dependent on flux density, which in turn is dependent on the frequency and voltage of the applied signal. In the next chapter we will describe methods of approximating the change to these elements over frequency. In this section we will explore the effects of extreme excitation on the inductance and core loss values.

The permeability of an inductor's core material indicates how much inductance is present. The formula describing the inductance of a solenoid is:

$$L = \frac{0.4 \pi N^2 A_c \mu_e}{l_m 10^8}$$

where:
N is the number of turns around the inductor's core
μ_e is the effective permeability of the core material
A_C is the effective magnetic area of the core
l_m is the magnetic path length of the core

Permeability, however, is a function of the magnetic flux density and for most materials used in the construction of transformers and inductors, looks something like the graph in Figure 9.

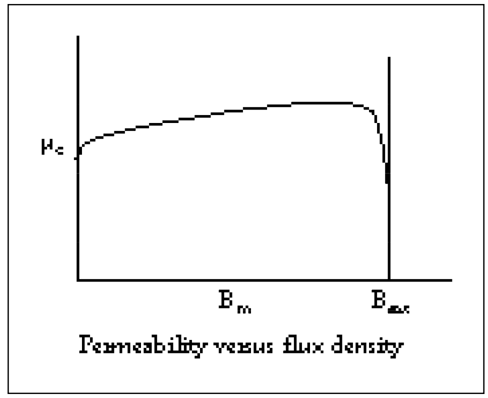


Figure 9

We can see that permeability, and hence inductance, rises as we approach the saturation flux density, B_{sat}, but drops abruptly as we reach saturation. Flux density is a measure of the magnetic energy in the core of the inductive device. The formula for flux density is:

$$B_M = \frac{V \times 10^8}{K_f N A f}$$

where:
B_M is the magnetic flux density in gauss
V is the voltage applied to the coil
N is the number of turns of the coil
A_C is the core area of the coil enclosed by the turns
f is the frequency of the voltage applied to the coil
K_f is a proportionality constant describing the energy in the wave form. (K_f=4.44 for sinusoids, K_f=4.0 for squarewaves)

We can see from this equation that flux density is proportional to applied voltage, but *inversely* proportional to the frequency of the applied voltage. This means that if a coil is operating at or near its saturation point, we can move its operating point out of saturation by either reducing the applied voltage or raising the frequency of that voltage. Many times reducing voltage is not feasible, so an increase in frequency is the next best option. This is the principle that leads to high-frequency power conversion and the efficiencies it can yield. We will return to this topic when we discuss *switchmode power conversion.*

Material	Saturation Flux Density (B_{sat})
Ferrite *High permeability*	3500 gauss
Ferrite *Power*	4500 gauss
Sheet steel *High nickel content*	7500 gauss
Sheet steel *Medium nickel content*	10000 gauss
Sheet steel *Silicon-based*	19000 gauss

Table 1

Saturation flux density limits are material dependent. Table 1compares the approximate saturation flux density of various materials.

Measurement	1	**2**	**3**	4	5	6	7	8	9
V (volts)=	1	**5**	**1**	1.1	1.25	1.35	1.4	1.45	1.5
Kf=	4.44	4.44	4.44	4.44	4.44	4.44	4.44	4.44	4.44
N (turns)=	450	450	450	450	450	450	450	450	450
Ac (cm2)=	0.103	0.103	0.103	0.103	0.103	0.103	0.103	0.103	0.103
f (Hz)=	1000	**1000**	**200**	200	200	200	200	200	200
Bm (gauss)=	485.9	2429.6	2429.6	2672.6	3037.0	3280.0	3401.5	3522.9	3644.4
L (henry)=	1.473	1.496	1.486	1.384	1.195	0.997	0.835	0.587	0.408

Table 2

The effects of saturation are illustrated in Table 2, where the inductance of an ferrite-core inductor is measured and recorded at several combinations of frequency and voltage.

From this we see that inductance remains reasonably constant at just under 1.5H for flux densities under about 2500 gauss. To see how permeability is reasonably independent of the combination of volts and hertz (assuming the ratio of the two is constant), note that measurements 2 and 3 have equivalent flux densities of 2429.6 gauss and roughly the same inductance—hence the same permeability—even though measurement 2 was made at 5V and 1000 Hz while measurement 3 was made at 1V and 200 hertz.

As the voltage is raised from 1V to 1.5V we see the effect of flux saturation as inductance drops off from 1.486 H to under 0.5 H.

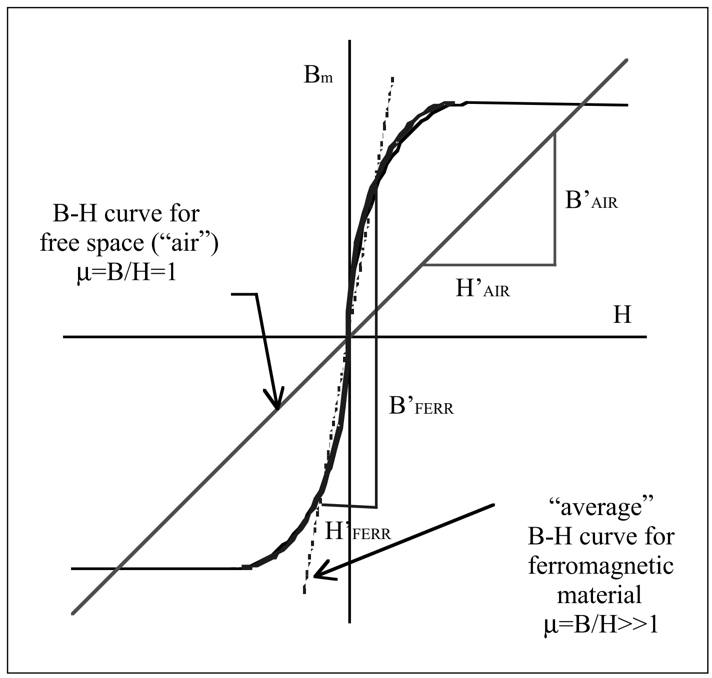


Figure 10

Permeability drops off significantly at the saturation point. We define permeability as the change in flux, B divided by the change in coercivity, H.

The absolute permeability of free space or "air" is $4\pi \times 10^{-7}$ henries per meter. We usually don't deal in absolute permeability when discussing ferromagnetic materials, so we consider this absolute permeability as a baseline, then refer our material permeability to it. We call this reference permeability the *relative permeability*, and by definition, the relative permeability of free space is 1.0. This is similar to the common use of the Celsius scale, which has reference points at the freezing and boiling points of water instead of the Kelvin scale which has a single reference point of absolute zero. When magnetics engineers speak of permeability, they are almost always referring to relative permeability, not absolute permeability.

Referring to Figure 10, you can see that the relative permeability of free space has a slope (rise/run) of 1.0. By contrast the curve traced by the operation of the core over the thick portion of the ferromagnetic material's curve has an average permeability many times greater than 1.0. The curve also shows the saturation points where further increase in H yields no change in B.

If you're wondering about the relationship between B, μ and H, you can think of the flux density B as being dependent on the applied voltage and independent of the permeability whereas H, the coercive force in the core, is a function of the core's permeability times the flux density in the core.

We will return to the B-H curve when we discuss the effects of saturation on signal distortion.

Chapter 2 The 'Complete' Equivalent Circuit

Now that we have defined the equivalent circuit of the inductor, we are ready to expand our model to encompass the entire transformer. Midcom uses a model developed by Thomas R. O'Meara[1] with modifications to account for the non-linear response of core loss resistance and magnetizing inductance. The "complete" model is shown in Figure 11.

We refer to this as the "complete" equivalent circuit of the transformer because it contains the minimum necessary elements to describe the behavior of well-designed transformers over their passbands. Other equivalent circuits may be used to describe special cases, but this circuit describes the bulk of all transformers and their usage. As Mr. O'Meara succinctly stated in his treatise: "It is not an easy matter to choose an equivalent circuit which is sufficiently complex to represent the physical transformer with reasonable accuracy, and yet is sufficiently simple to permit ready analysis or synthesis."

Frequency response (amplitude distortion)

From the previous chapter we see that the equivalent circuit of an inductor forms three of the elements of the complete transformer equivalent circuit: L_{pri}, R_{core} and R_{Pri}. The inductor part of that triad, which we call the magnetizing inductance, is largely responsible for the cutoff point f_L, of the low-frequency response. Similarly, the high-frequency cutoff point f_H, is dependent on a different set of three elements: C_{pri}, L_{leak} and C_{sec}.

Transformers having two or more decades of bandwidth (f_H>100 f_L) can be thought of as having simpler and distinctly different equivalent circuits below and above the mid-point frequency of their passbands, typically the geometric mean of their low- and high- cutoff frequencies, $\sqrt{f_L \bullet f_H}$.

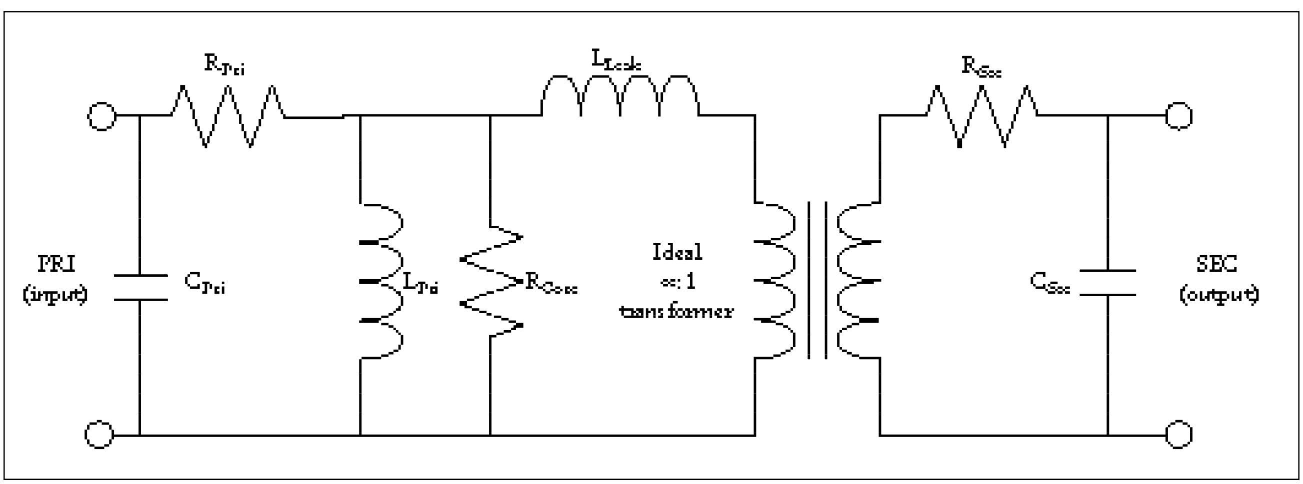


Figure 11: The 'complete' transformer equivalent circuit

[1] T.R. O'Meara, "Analysis and Synthesis with the 'Complete' Equivalent Circuit for the Wide-Band Transformer" *AIEE Transactions*, March, 1962

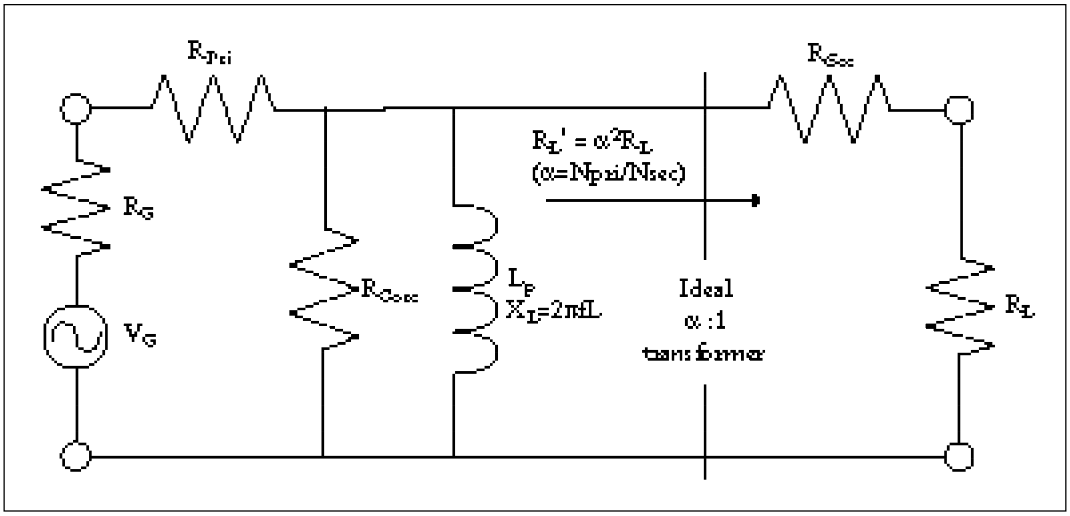


Figure 12: Low-frequency ($f<f_c$) equivalent circuit

Low-frequency analysis

By inspecting the low-frequency model we can see that the primary *magnetizing inductance*, L_P, is the factor which controls the low-frequency cutoff point.

A finite amount of primary inductance will result in less-than-perfect frequency response, return loss and impedance shift. While we could analyze the entire equivalent circuit to determine the minimum inductance required to meet a given specification, we can apply a shortcut method that provides reasonably accurate results.

Calculating the approximate minimum inductance to meet a given frequency response

Let G represent the attenuation of a voltage in decibels (dB) at a low frequency, f. G must be a positive value since we're assuming the roll-off already represents a signal loss. Then:

$$A = 10^{\left(\frac{G}{20}\right)}$$

$$X_{L\,\min} = \frac{\left(\dfrac{R_G R_L'}{R_G + R_L'}\right)}{\sqrt{A^2 - 1}} = \frac{R_G}{2\sqrt{A^2 - 1}}$$

(when source is matched to the transformed load, i.e. $R_G=R_{L'}$) and

$$L_{Min} = \frac{X_{L\min}}{2\pi f}$$

where:

X_{Lmin} is the minimum reactance required to meet attenuation G at frequency f
R_G is the generator (source) resistance
R_L is the load resistance
R_L' equals $\alpha^2 R_L$, which is the load resistance compensated for the turns ratio, α

EXAMPLE: (Source matched to load via a matching transformer) Find the minimum primary magnetizing inductance required to support a frequency response of -1 dB maximum roll-off at 100 Hz given a 50 ohm generator impedance, a 120 ohm load impedance with the transformer (lossless) having a primary-to-secondary turns ratio of 0.6455 (value chosen to provide proper match between 120 ohms and 50 ohms):

$$A = 10^{\left(\frac{G}{20}\right)} = 10^{\frac{1}{20}} = 1.122$$

$$R_L' = \alpha^2 R_L = (0.6455)^2 (120) = 50.0\Omega\,,$$

and since $R_G=R_L'$

$$X_{L\,\min} = \frac{\left(\dfrac{R_G R_L'}{R_G + R_L'}\right)}{\sqrt{A^2 - 1}} = \frac{R_G}{2\sqrt{A^2 - 1}} = \frac{50}{2\sqrt{(1.122)^2 - 1}} = 49.13\Omega$$

$$L_{\min} = \frac{X_{L\min}}{2\pi f} = \frac{49.13}{2\pi 100} = 78.2 \text{ mH}$$

EXAMPLE: (Source and load mismatched as a result of incorrect turns ratio) Find the inductance required to support a frequency response of ±1 dB at 100 Hz given a 50 ohm generator impedance, a 120 ohm load impedance and a lossless ***1:1*** *transformer between the two:*

$$R_L' = \alpha^2 R_L = (1)^2 (120) = 120\Omega$$

$$X_{L\,\min} = \frac{\left(\dfrac{R_G R_L'}{R_G + R_L'}\right)}{\sqrt{A^2 - 1}} = \frac{\left(\dfrac{(50)(120)}{50 + 120}\right)}{\sqrt{(1.122)^2 - 1}} = 69.37\Omega$$

$$L_{Min} = \frac{X_{L\min}}{2\pi f} = \frac{69.37}{2\pi 100} = 110.4 \text{ mH}$$

From this we can see that more inductance is required to support the 120 ohm impedance.

Impedance shift

A transformer will introduce an impedance shift into a transmission system as a result of its finite magnetizing inductance. The amount of shift is directly related to the amount of magnetizing inductance: more inductance means less shift. Since it is desired that a transformer match its driving impedance as closely as possible, it becomes an important matter to control impedance shift to keep the mismatch within specified limits. Once the allowable shift is known, the minimum required inductance to support the shift may be found. A shortcut approach is shown here will provide results accurate enough for an initial design.

Determining the minimum magnetizing inductance required to support a given impedance shift

If we let ∂ represent the impedance shift allowed, where

$$\delta = \frac{Z_{nom} - Z_{\min}}{Z_{nom}} \quad \text{and } \partial \text{ x } 100 = \text{percent allowable shift,}$$

then

$$X_L = Z_0 \frac{(1-\delta)}{\sqrt{1-(1-\delta)^2}} \quad \text{and} \quad L_{\min} = \frac{X_L}{2\pi f}$$

where:
Z_0 is the reference impedance against which the impedance shift is to be compared
X_L is the minimum inductive reactance required to support the impedance shift
L_{min} is the minimum inductance required to provide the inductive reactance
f is the frequency at which the impedance shift is to be determined

EXAMPLE: A 600 ohm network is to be connected to a transformer such that the impedance shift caused by the transformer must be less than 20% at 300 Hz.

$$\delta = \frac{\text{percent shift}}{100} = \frac{20\%}{100\%} = 0.2$$

$$X_L = Z_0 \frac{(1-\delta)}{\sqrt{1-(1-\delta)^2}} = (600)\frac{(1-0.2)}{\sqrt{1-(1-0.2)^2}} = 800\Omega$$

$$L_{\min} = \frac{X_L}{2\pi f} = \frac{800}{2\pi 300} = 0.424H$$

Thus 0.424 henries will provide enough inductive reactance to prevent an impedance shift more than 20% below 600 ohms, or 480 ohms minimum.

Return loss

Since a transformer can shift the impedance of a network, it can also affect return loss. Return loss is defined as the ratio of a transmission system's reflected energy to incident energy expressed in terms of

[2] Consult the abscissa of a complete Smith chart for a means of comparing these terms.

decibels. Since the ratio of reflected-to-incident energy is the definition of reflection coefficient, return loss is effectively the expression of reflection coefficient expressed in decibels. Standing wave ratio (SWR) is another closely related to return loss.[2]

Good return loss is important in a communications system because reflections of the incident wave may interfere with a signal traveling in the same direction as the reflection. These reflections are perceived as echoes and can cause complete breakdown of the communications process. While modern communication systems may employ sophisticated means of suppressing or canceling echoes, it is better to prevent them from occurring in the first place. Paying proper attention to return loss is a good way to assure this.

Since the term return *loss* signifies a relative decrease in the signal, we at Midcom have taken the stance that the 'minus' sign is implied by the term *loss*. Thus an echo which is 20 dB down from the incident wave is said to represent a return loss of 20 dB, *not* -20 dB. The difference makes itself evident in the two of the formulae shown here where the "Midcom" definitions are calculated with the reciprocal of the arguments to make the sign come out correctly.

Return loss may be calculated from the impedances of two elements in a communications network. The formulae for reflection coefficient, SWR and return loss are:

Reflection coefficient: $$\Gamma = \frac{Z_O - Z_M}{Z_O + Z_M}$$

Standing Wave Ratio (SWR): $$s = \frac{1+|\Gamma|}{1-|\Gamma|}$$

"Midcom" Return Loss: $$r = 20\log_{10}\left[\Gamma^{-1}\right] = 20\log_{10}\left[\frac{Z_O + Z_M}{Z_O - Z_M}\right]$$

"Midcom" Return Loss, using real and reactive elements:

$$r = 20\log_{10}\left[\frac{(R_O + R_M)^2 + (X_O + X_M)^2}{(R_O - R_M)^2 + (X_O - X_M)^2}\right]^{1/2}$$

where:
Z_0 *is the reference impedance of a given network*
Z_M *is the impedance of a device to be measured against the reference impedance*
R_0 *is the resistive portion of the reference impedance*
R_M *is the resistive portion of the measured device's impedance*
X_O *is the reactive portion of the reference impedance*
X_M *is the reactive portion of the measured device's impedance*

EXAMPLE: Find the return loss at 1000 Hz of a device whose impedance is 484 + j29 ohms versus a reference network consisting of a 600 ohms in series with a 2.16μF capacitor.

First, calculate the reactance of the 2.16μF capacitor at 1000 Hz.

$$X_0 = X_C = \frac{-1}{2\pi f C} = \frac{-1}{2\pi(1000)(2.16 \text{ x } 10^{-6})} = -73.68\Omega$$

$$r = 20\log_{10}\left[\left[\frac{(R_O + R_M)^2 + (X_O + X_M)^2}{(R_O - R_M)^2 + (X_O - X_M)^2}\right]\right]^{1/2} = 20\log_{10}\left[\left[\frac{(600+484)^2 + (-73.68+29)^2}{(600-484)^2 - (-73.68-29)^2}\right]\right]^{1/2}$$

$$r = 16.9 \text{ dB}$$

Determining the minimum inductance require to meet a given return loss

As we did with impedance shift, we can calculate a minimum inductance required to meet a given return loss. A shortcut approach that assumes the transformer is well-matched at mid-band and has a primary winding resistance less than 5% of the reference impedance is shown here:

$$X_{L\min} = \frac{R_0}{2}\left(10^{\left(\frac{r}{10}\right)} - 1\right)^{1/2}$$

then $$L_{\min} = \frac{X_L}{2\pi f}$$

EXAMPLE: Find the inductance required for a transformer to provide 14 dB return loss at 200 Hz versus a 600 ohm resistive reference network. Assume the resistance of the primary winding is less than 30 ohms.

$$X_L = \frac{R_0}{2}\left(10^{\left(\frac{r}{10}\right)} - 1\right)^{1/2} = \frac{600}{2}\left(10^{\left(\frac{14}{10}\right)} - 1\right)^{1/2}$$

$$X_L = 1473\Omega$$

$$L_{\min} = \frac{X_L}{2\pi f} = \frac{1473}{2\pi 200}$$

$$L_{\min} = 1.172 \text{ H}$$

If the primary winding's resistance is greater than 5% of the reference impedance, it is still possible to estimate the inductance required. Since the primary winding resistance helps to dissipate the energy of the incident wave, we can apply a correction factor to the initial return loss to account for the loss due to the winding resistance[3]. In this case, use:

[3] To demonstrate the principle that high resistance can promote better return loss when the reference network is purely resistive, recalculate the first example assuming the primary resistance is 100 ohms instead of 30. You should find that the higher resistance provides about 3 dB additional return loss, lessening the minimum inductance required from 1.172 to 0.8 henries.

$$r' = r - \left(10A^2 + 17A\right)$$

where:

r' is the return loss target, compensated for winding resistance
A is the ratio of primary winding resistance to reference resistance, $A = \dfrac{R_{pri}}{R_0}$
The correction factor is accurate for r > 16 dB and A < 0.25.

EXAMPLE: Find the inductance required to support 10 dB return loss versus 600 ohms at 300 Hz assuming the resistance of the primary winding is 65 ohms.

Calculate the ratio of primary resistance to reference resistance, A:

$$A = \frac{R_{pri}}{R_0} = \frac{65}{600} = 0.1083$$

Calculate the corrected return loss target value:

$$r' = r - \left(10A^2 + 17A\right) = 10 - \left(10(0.1083)^2 + 17(0.1083)\right) = 8.04\,dB$$

$$X_L = \frac{R_0}{2}\left(10^{\left(\frac{r'}{10}\right)} - 1\right)^{1/2} = \frac{600}{2}\left(10^{\frac{8.04}{10}} - 1\right)^{1/2}$$

$$X_L = 695\Omega$$

$$L_{\min} = \frac{X_L}{2\pi f} = \frac{695}{2\pi 300}$$

$$L_{\min} = 0.369H$$

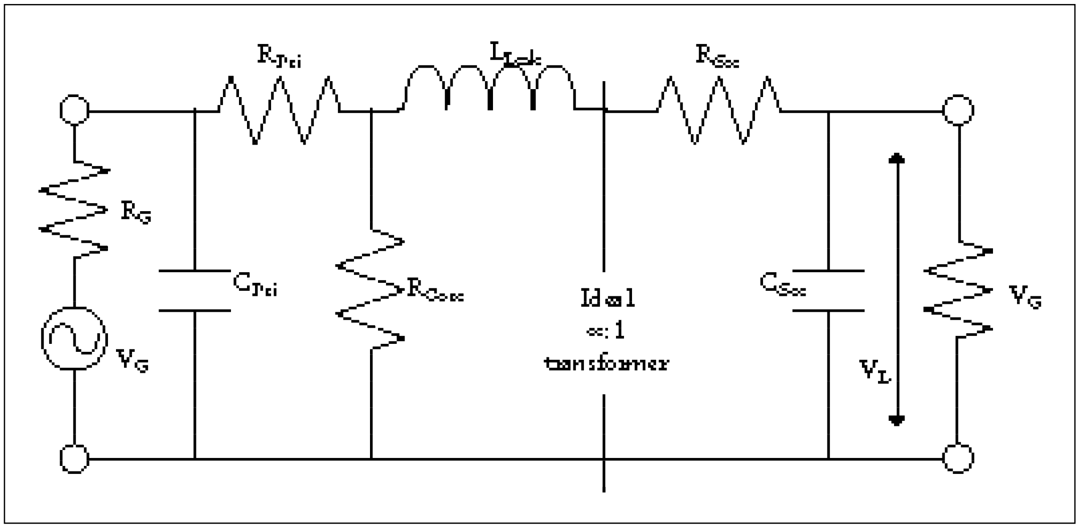


Figure 13

High-frequency analysis

In the high-frequency simplified model, the high-frequency cutoff point is controlled by the leakage inductance and the primary and secondary distributed capacitances, as in Figure 13.

While there are no shortcut methods for determining limits for leakage inductance and distributed capacitances, a quick review of the analysis of the high-frequency model will help convey the relationships between those elements and the source and load impedances. For now, we will consider only the effects due to leakage inductance, as this it is generally the most significant factor to cause high-end roll-off.

Calculating high-frequency response due to leakage inductance

Using the voltage division property of a series circuit we see that the output, V_L as a function of the generator voltage, V_G, which we will call A, is:

$$A = \frac{V_L}{V_G} = \frac{R_L}{R_L + R_G + jX_{Lleak}}$$

Finding A gives us the overall loss circuit "gain", but we must subtract the losses due to the source and load to find the transformer loss, TL. It becomes easier to manage if we express these losses in decibels:

$$TL = 20\log(|A|) - 20\log\left(\frac{R_L}{R_L + R_G}\right)$$

Phase response

The phase response of the transformer, like its frequency response, is a function of the reactive elements as described in the equivalent circuit. The low-frequency phase response of the transformer is largely dependent on its magnetizing inductance while the high-frequency response is influenced by its leakage inductance and distributed capacitances. Thankfully, the phase response is generally of secondary import for most well-designed transformers, assuming the amplitude response does not swing wildly about. The phase response and group delay characteristics of a transformer are almost always less troublesome to the circuit designer than the characteristics of the transmission medium. If the phase angle is a linear function of frequency, the delay will be constant and not pose problems to the designer. For most simply-constructed transformers, the phase angle is relatively well-behaved within its intended passband and should pose no great concern to the circuit designer.

The phase response based solely on leakage inductance, using the preceding definitions is:

$$\theta = TAN^{-1}\left(\frac{\mathrm{Im}(A)}{\mathrm{Re}(A)}\right)$$

Insertion loss and transducer loss

The two terms *insertion loss* and *transducer loss* seem to suffer from confusion in the same way as the words *uninterested* and *disinterested*. The terms are related, but different[4] . Things get even more confusing when, under certain conditions, the two are effectively identical. One article that does an excellent job describing the differences in detail was published in 1980[5] We will cover the gist of the information without delving into the derivations of the formulae. The definitions of the terms are listed here for reference. You may substitute "telephone facility" or "lossy device" for "transformer" if you wish to make the explanations generic.

Note that if R_g=R_L, the second term of the "insertion loss" definition equates to zero. Thus when the generator and the load are matched, transducer loss and insertion loss are the same. Remember that the Maximum Power Transfer theorem states that maximum power is transferred from source to load when the impedance of the generator is equal to the impedance of the load. If you measure the power delivered to a load with a transformer in place, then measure the power delivered to a load with the transformer bypassed, you won't be taking into account the impedance match provided by the transformer.
If instead, you measure the maximum power available from the generator (which assures a matched condition), then measure the power delivered to the load with

$$\text{Transducer loss} = \frac{\text{Maximum power available from the source}}{\text{Power delivered to the load with the transformer inserted}} = 10\log_{10}\left[\frac{(V_g)^2 R_L}{4R_g(V_L)^2}\right] dB$$

$$\text{Insertion loss} = \frac{\text{Power delivered to the load with the transformer bypassed}}{\text{Power delivered to the load with the transformer inserted}} = \text{Transducer loss} - 10\log_{10}\left[\frac{(R_L+R_g)^2}{4R_gR_L}\right] dB$$

Figure 14

[4] For the record, *disinterested* is what we expect our judges to be when considering two sides of a legal dispute. *Uninterested* is what we hope the police to be when we note a positive discrepancy between our vehicle speed our speedometer.

[5] "Measured Loss is NOT Insertion Loss" by Richard M. Hardy, Telephone Engineer and Management, March 15, 1980. Also available as Midcom Technical Note #15.

the transformer inserted, you will measure the true losses of the transformer, regardless of losses due to impedance mismatch.

$$\text{Transducer loss} = 10\log_{10}\left[\frac{(V_G)^2 R_L}{4R_G (V_L)^2}\right] dB = 10\log_{10}\left[\frac{(1.85)^2(900)}{4(600)(0.95)^2}\right] dB = 1.529\ dB$$

$$\text{Insertion loss} = \text{Transducer loss} - 10\log_{10}\left[\frac{(R_L + R_G)^2}{4R_G R_L}\right] dB$$

$$= 1.529 - 10\log_{10}\left[\frac{(900+600)^2}{4(600)(900)}\right] = 1.529 - 0.177 = 1.352\ \text{dB}$$

Important note: To actually measure transducer loss, you must be able to measure the actual generator voltage 'behind' the source impedance. Thus we define V_G as the voltage between R_G, an externally-supplied generator impedance, and R'_G, the built-in impedance of our real-world generator. The full circuit is shown here for clarification:

An interesting point about the circuit in Figure 15 is to note that the internal impedance of the generator, R_G', is completely irrelevant to the measurement of either insertion loss or transducer loss, as long as V_G is constant. Of course holding V_G constant has the same effect as taking a measurement with a perfect voltage source: one with zero source impedance. In practical laboratory measurements, it is a good idea to use an external R_G anyway, as it is much easier to verify the impedance of a stand-alone resistor than the internal impedance of a particular generator. Use of an external R_G also provides more flexibility and in fact will allow use of virtually any generator, whether it be 50 ohms, 600 ohms or 10 megohms, to make loss measurements.

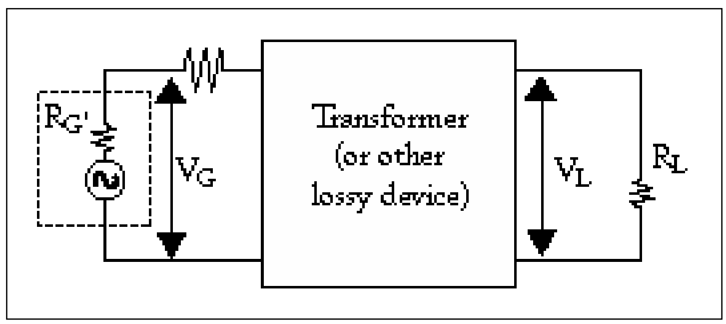


Figure 15

EXAMPLE: Find the insertion loss and transducer loss of a transformer designed to match a 600 ohm source to a 900 ohm load. Assume a 600 ohm generator provides 1.85 volts at its output with the transformer delivering 0.95 volts to its 900 ohm load:

The second term, equal to 0.177 dB in our example, effectively describes the lack of maximum power transfer as a result of the impedance mismatch.

Aside from the fact that insertion loss and transducer loss are equivalent when R_G=R_L, another reason people mistakenly measure insertion loss instead of transducer loss is that insertion loss is a bit easier to measure. What they may not know is that the conversion from insertion loss to transducer loss is fairly simple and involves only the introduction of the second term describing the effects of the mismatch. In an automated test environment it becomes a simple matter to measure V_G and V_L to calculate the insertion loss, then apply the correction factor based on R_G and R_L. This is covered in greater detail in Midcom Technical Note #16, "Measuring Transducer Loss using a Network/Spectrum Analyzer". It is also available at the Midcom web site, http://www.midcom-inc.com/technotes/tn16.pdf (PDF format).

Nonlinear elements and their effect on the equivalent circuit

Much to the consternation of PSPICE users, transformers contain frequency-dependent modeling elements. Two elements that change significantly with frequency are the magnetizing inductance and the core loss resistance[6]. Both of these elements change due to variation in applied flux density. Inductance is a function of permeability, which may vary greatly (100:1 or more) depending on flux density. Midcom engineers characterized the frequency-dependency of magnetizing inductance and core loss in the early 1980s. These approximations assume the transformer core is not operating at or near its saturation flux density. The formulae for the two elements are shown here:

Inductance change as a function of frequency

$$L' = L(f_R)\left[\frac{f}{f_R}\right]^{-\left(\frac{\ln(\alpha_L)}{\ln\left(\frac{f_R}{f_L}\right)}\right)}$$

where:
f is the frequency at which we need to determine the inductance
L' is the inductance at a given frequency, f
L(f_R) is the inductance at a reference frequency, f_R
f_R is a reference frequency, usually chosen to be near the transformers mid-band frequency
f_L is a frequency lower than f_R, usually about one-fifth the value of f_R
a_L is ratio of inductances at a low frequency, f_L to the reference frequency, f_R, or $\alpha_L = \frac{L(f_L)}{L(f_R)}$

For audio frequency transformers, Midcom uses f_r=1000 Hz and f_L = 200 Hz. Most audio frequency transformers have α_L values ranging from 1.0 to about 2.5. This means that the inductance at 200 Hz may range between 1.0 and 2.5 times the inductance at 1000 Hz. An increase in inductance is beneficial to our efforts to meet low-frequency response and return loss requirements. It may cause PSPICE and other simulations to provide inaccurate results unless special precautions are taken to scale the inductance over frequency as described above.

EXAMPLE: Find the inductance at 697 Hz of an audio transformer having an inductance of 1.5 H at 1 kHz and 2.25 H at 200 Hz.

$$\alpha_L = \frac{L(f_L)}{L(f_R)} = \frac{2.25}{1.25} = 1.8$$

$$L' = L(f_R)\left[\frac{f}{f_R}\right]^{\frac{\ln(\alpha_L)}{\ln\left(\frac{f_R}{f_L}\right)}} = 1.25H\left[\frac{697Hz}{1000Hz}\right]^{-\left(\frac{\ln(1.8)}{\ln\left(\frac{1kHz}{200Hz}\right)}\right)} = 1.43H$$

Core loss change as a function of frequency

$$R_C' = R_C(f_R)\left[\frac{f}{f_R}\right]^{\alpha_{Rc}}$$

[6] Winding resistance also changes with frequency due to skin effect.

where:

f is the frequency at which we need to determine the core loss resistance
R_C' is the inductance at a given frequency, f
$R_C(f_R)$ is the core loss resistance at a reference frequency, f_R
f_R is a reference frequency, usually chosen to be near the transformers mid-band frequency
a_{Rc} is the core loss resistance factor, which usually ranges from 0.35 to 0.45

EXAMPLE: Find the effective core loss resistance at 3400 Hz for a voiceband transformer having 10K ohms of core loss resistance at 1kHz and a core loss resistance factor of 0.35:

$$R_C' = R_C\left(f_R\right)\left[\frac{f}{f_R}\right]^{\alpha_{Rc}} = 10k\Omega\left[\frac{3400}{1000}\right]^{0.35} = 15347\Omega$$

Neither of these relationships takes into account variations due to signal level. Nor do they take into account effects due to temperature or other external environmental effects. Effects due to temperature are discussed in a later chapter.

The time-domain response

Until now we have concerned ourselves only with the transformer's response in the frequency domain. We will now take a look at its response in the time domain. This will also serve as a basis for the chapter dealing with switchmode power applications.

A pulse applied to the primary winding of an ideal transformer would provide an exact replica of the pulse's waveform at its unloaded secondary, scaled in amplitude by the transformer's turns ratio. In practice, a transformer exerts a number of distortions on the pulse which in extreme cases may cause the resultant pulse to be unusable.

When we were dealing in the frequency domain, we found it useful to break the transformer equivalent circuit into two models: one representing the low-frequency response, the other representing the high frequency response. The same applies to the time domain except we refer to the models as describing the "top period" and the "edge periods".

The "top" period

The top period refers to the duration just after the pulse is applied to the transformer until just before the pulse is removed. Figure 15 shows a simplified view of the various periods of a pulse.

The *droop*, D, and backswing, B, of a transformer are functions of its finite magnetizing inductance. By

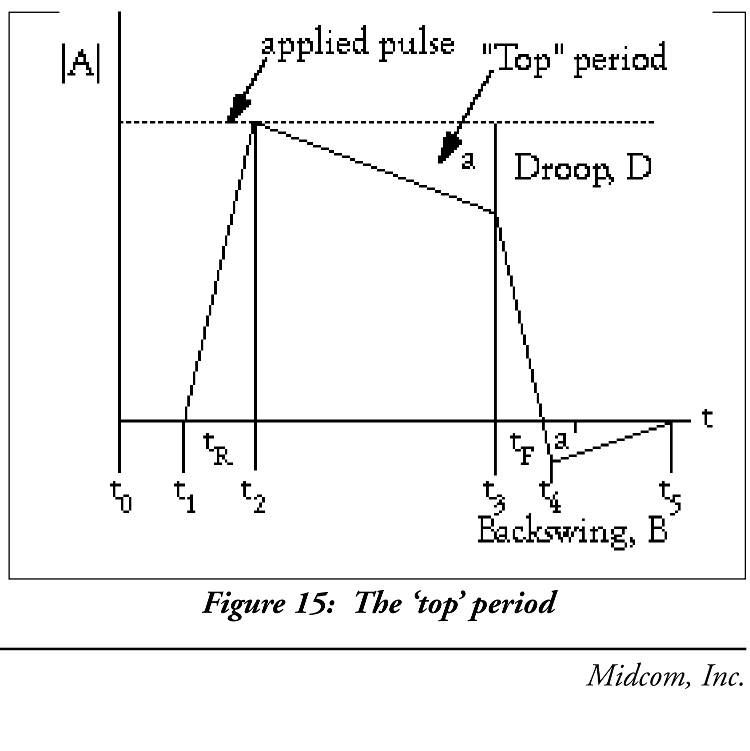


Figure 15: The 'top' period

inspecting the low-frequency equivalent circuit we can see how a finite magnetizing inductance is the cause of droop and backswing. (See Figure 16.)

Droop is caused *by the increase in magnetizing current that occurs as the magnetic flux builds around Lp. The current through Lp as a function of time is expressed by:*

$$i_{L_P}(t) = \frac{V}{R}\left(1 - e^{\frac{Rt}{Lp}}\right)$$

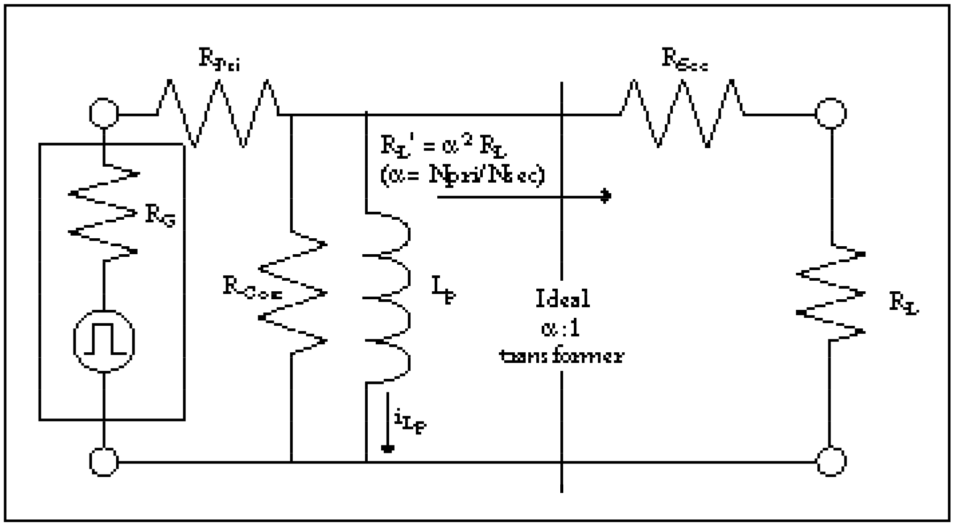


Figure 16

From this we can see that initially, at t=0, no current flows and the voltage appearing at the secondary is purely a function of the voltages applied to the primary, resistances in the circuit and the turns ratio of secondary to primary. At t>0, current begins to flow in L_P, causing a voltage drop to appear across it. When the applied pulse voltage returns to zero at time t_3, the inductor current continues to flow since current cannot change instantaneously through it. At time t_3, the inductor begins to act as a current source, pushing current in the same direction it was flowing just prior to t_3. Since the source is at zero volts, a voltage of the opposite polarity (negative-going) is developed across the "ideal n:1 transformer". This voltage is scaled by the turns ratio and is seen as the backswing of the waveform shown in figure 5. The *equal area theorem* states that the energy missing as a result of droop is returned to the circuit during the backswing period. See time period t_4-t_5 in Figure 15.

Droop may be approximated (neglecting effects due to winding resistance) as follows:

$$D = 100\frac{R_G R_L'}{L_p\left(R_G + R_L'\right)}T$$

where:
D is droop in percent
L_P is the magnetizing inductance of the primary in henries
R_G is the impedance of the generator in ohms
R'_L is the effective load impedance, scaled by the primary-to-secondary turns ratio: $R_L' = \alpha^2 R_L$
T is time in seconds

A more useful rearrangement of the formula allows us to find a minimum value of Lpri for a given droop:

$$L_P = 100\frac{R_G R_L'}{D\left(R_G + R_L'\right)}T$$

EXAMPLE: Given a circuit with a 6Vpeak generator with a source resistance of 25 ohms, a transformer with turns ratio 1:2 (pri:sec) driving a 100 ohm load impedance, find the minimum inductance required to support a worst-case pulse droop of 1.2V after 162nS:

$$D = \frac{1.2V}{6V} = 0.2 = 20\%$$

$$R_L' = \alpha^2 R_L = (0.5)^2 (100) = 25\Omega$$

$$L_P = 100 \frac{R_G R_L'}{D(R_G + R_L')} T = 100 \frac{(25)(25)}{20(25+25)} 162nS = 10.13\ \mu\text{H}$$

The "edge" periods

The edge periods consist of rise time, fall time and ringing, which encroaches the top period if it is extreme. We may refer to the high-frequency equivalent circuit for our analysis of the edge periods as shown in Figure 17.

The transformer's rise time may be calculated from the distributed capacitances and leakage inductance. It is assumed that the rise time is the period defined when the output pulse is between 10% and 90% of its full output level. A first-order approximation of rise time is:

$$t_R = 1.52\sqrt{L_{Leak} C_{Dist}}$$

where:
t_R is the rise time in seconds from the 10% to 90% points of the pulse output
L_{Leak} is the leakage inductance appearing at the primary
C_{Dist} is a lumped-parameter equivalent of the capacitances shown as C_{pri} and C_{sec} in figure 6.

The fall time of the pulse, t_F is the same duration as the rise time, assuming there are no nonlinear elements involved in the circuit.

Ringing of the output waveform occurs when the load and source are mismatch with the load impedance being much higher than the appropriate value to meet the matched condition (underdamped condition)

Symanski discusses a means of calculating ringing

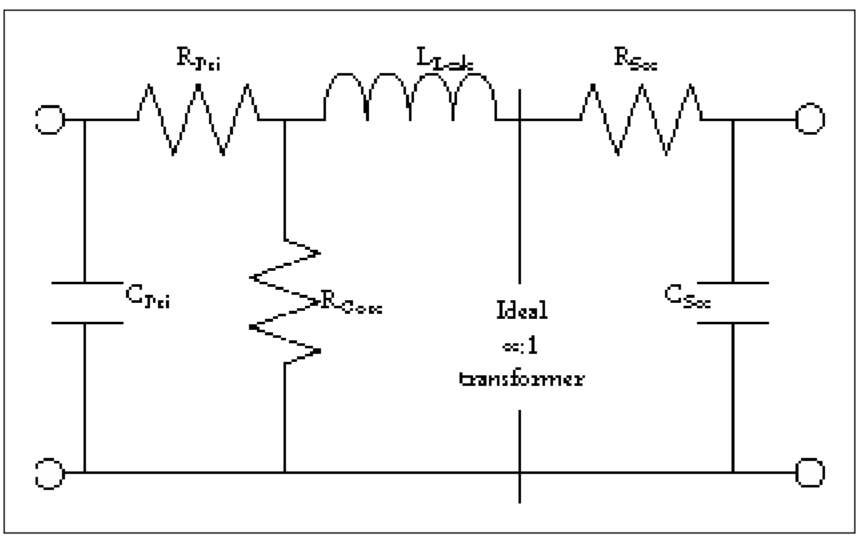


Figure 17

frequency and the decay of the ringing waveform in his treatise on pulse transformer design.[7] Overshoot is also discussed in that article.

Effects of saturation and the "E-T" constant

Saturation of the magnetic core will cause distortion of the pulse shape. The most pronounced effect is that the droop will be much greater than an unsaturated transformer since saturation causes a significant drop in inductance. Figure 18 shows the effects of saturation on pulse shape.

It is convenient to describe the transformer's capability in terms of its voltage-time constant, also known as its ET constant. The ET constant effectively describes the "area" of a pulse where the pulse height and width are measured in volts and seconds. Once a transformer's ET constant is known, it is possible for it to support any result of the voltage-time product up to the maximum rated ET constant. This is really no different from flux density expressed in the time domain where time is simply the reciprocal of the frequency of the repetitive pulse applied to the transformer.

EXAMPLE: If a transformer can support a squarewave pulse of 5Vpeak maximum at 10kHz, find its ET constant:

$$T = \frac{1}{f} = \frac{1}{10kHz} = 0.1mS$$

$$K_{ET} = (E)(T) = (5V)(0.1mS) = 0.5mVS$$

Can the transformer described above support at 12V, 25uS pulse?

$$K_{ET}(desired) = (12V)(25uS) = 0.3mVS$$

So, yes, this transformer can support the 12V, 25uS pulse.

EXAMPLE: What is the maximum "on" time for the above transformer given a 3.3Vpeak pulse?

Rearranging the equation

$$K_{ET} = 0.5mVS$$

to yield

$$T = \frac{K_{ET}}{E} = \frac{0.5mSV}{3.3V} = 0.152mS$$

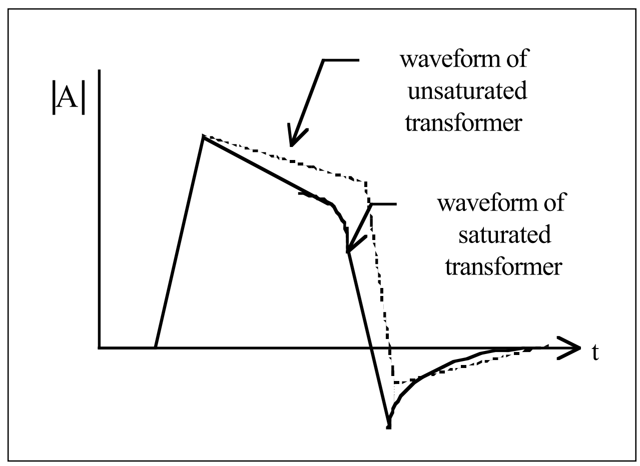


Figure 18

Coefficient of coupling versus leakage inductance

As discussed at the beginning of this chapter, it is possible to describe the equivalent circuit of a transformer in several different ways. One method which finds popularity in electrical engineering texts uses the coefficient of mutual inductance, M, and coefficient of coupling, k, to replace the turns ratio and leakage inductance used in our "complete" model. Mutual inductance, as the name implies, describes the concept that transformer coupling is bi-directional. In other words, a signal applied to the primary winding causes a related signal to appear at the secondary windinxg *and vice-versa*. In fact, a transformer may pass signals in both directions *simultaneously*. (Try *that* with an op-amp!)

[7] "A Unified Approach to Pulse Transformer Frequency and Pulse Response", *EEE*, October 1965, Eugene S. Szymanski

Mutual inductance is equal to the geometric mean of the primary and secondary inductances. Mutual inductance is described schematically as in Figure 19.

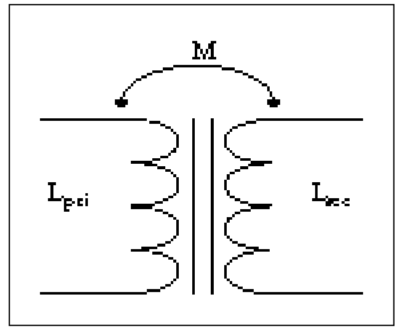


Figure 19

The mutual inductance, M, describes how much inductance is shared between the physical windings which ultimately results in the expression of the voltage step-up or step-down ratio. When coupling is perfect (only *theoretically* possible), the mutual inductance is equal to the geometric mean of the primary and secondary inductances. In practice, coupling is less than perfect and we employ the coupling coefficient, k, to describe the decoupling of the primary and secondary inductances:

$$M = k\sqrt{L_{PRI}L_{SEC}}$$

While there exist applications for transformers having k values less than a few tenths, most useful transformers have k values ranging from 0.8 to 0.999. Conversion between k and leakage inductance is possible if an accurate means of accurately determining magnetizing inductance and leakage inductance is available. The conversion is:

$$L_{\mathrm{Pr}\,iLeak} = L_{PRI}\,(1-k)^2$$

where $L_{PriLeak}$ is the leakage inductance measured at the primary

A similar conversion is possible when measuring the secondary side by:

$$L_{SecLeak} = L_{SEC}(1-k)^2$$

As stated earlier, an *accurate* means of determining the actual primary inductance must be made for this conversion to be meaningful, particularly when k approaches unity. Inaccurate measurements of L_{PRI} $L_{PriLeak}$ or $L_{SecLeak}$ may result in calculated values of k which exceed 1 which is not possible with real-world devices. For background information on modeling techniques, refer to Midcom Technical Note # 82 which may be downloaded from the Midcom web site,

http://www.midcom-inc.com/technotes/tn82.doc (Microsoft Word 6.0).

Chapter 3 Safety

Safety concerns are just about the number one reason why people use transformers. Transformers provide low-cost, passive and rugged barriers between protected and unprotected circuitry. We will delve only lightly into the specifics of the various global safety specifications, covering basic safety concepts and review only items which are unlikely to change as new standards introduced and existing standards are rewritten.

> *IMPORTANT: The contents of this chapter are not meant to convey the absolutes of safety requirements. To obtain current information on the status of various safety regulations you must contact the safety agency and obtain the latest official release of the relevant safety standard. You may also refer to Midcom Technical Note #79 which describes differences between a few common global telecom safety standards.*

While there are several safety specifications throughout the world, by and large they contain much the same sort of classifications although the terminology may be slightly different from one to the next. Since Midcom has a large stake in the telecommunications industry, we will often refer to the International Electrotechnical Commission's document, IEC950 or its various derivatives. IEC-950 is where EN60590 and UL1950 and CSA 950 (Canada) have found their roots. A closely related document is used in Australia under the number AS/NZS 3260. IEC-950 has had several amendments and as of this writing the most commonly-accepted issue is amendment 4.

We should first define a few terms. The terms *creepage* and *clearance* are used to describe the distances between conductors. Clearance is the line-of-sight distance between conductors while creepage is the shortest path along the surface of the insulating medium between the conductors.

Creepage and clearance

Creepage paths are subject to breakdown if the insulating medium between the conductors becomes contaminated. The presence of a DC voltage present on the conductors accelerates the rate of contamination and breakdown through a process known as *tracking*. Insulators are rated in terms of their CTI or *comparative tracking index*. Tracking refers to the formation of conducting paths made by the contaminants, typically carbon-based, that collect on an insulating medium. Materials are grouped by their ability to withstand the formation of carbon tracks. The highest

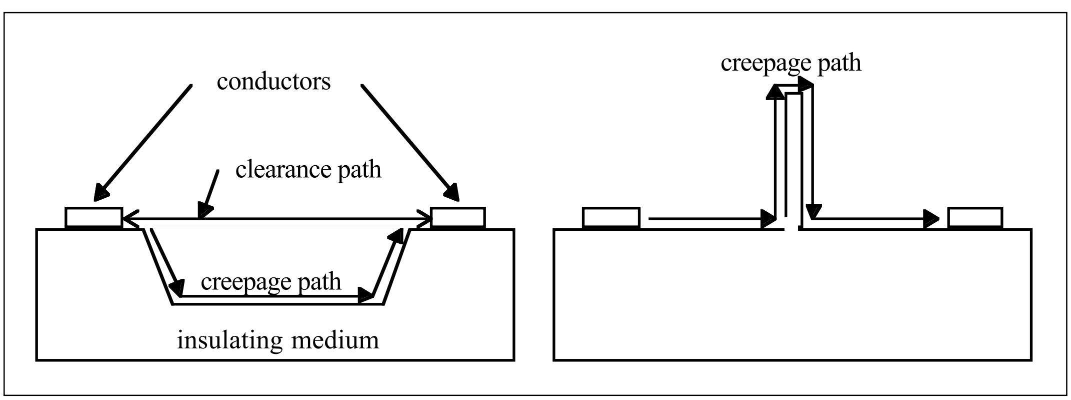


Figure 20: Creepage paths

voltage a test sample of material can be subjected to before it exhibits tracking determines the material's CTI rating group.

Two conditions must be present for tracking to occur:

1. A voltage difference between two conductors separated by an insulating medium, and
2. A contaminant that provides a charge carrier. Moisture accelerates the tracking process.

Materials may also be rated for their flame-retardancy, typically measured by the time it takes a sample of the material to self-extinguish after being lit on fire. Due to the nature of the filling materials used in insulating plastics, it is not uncommon that a given plastic may have an excellent flammability rating, but have a relatively poor CTI rating, although recent advances in material research are addressing this issue.

Creepage and clearance distances form part of the list of requirements that come into play when a safety agency, or one of its designated labs, reviews the issue of whether your device complies with that agency's safety standard. Other requirements include, but are not limited to, dielectric withstanding voltage, flammability of the materials used in the construction of the equipment, the environment in which the equipment is designed to operate, whether the device's operational enclosure is open or sealed and of course the power sources and other devices to which it may be connected. Even the product's marketing plan may be called into the approval review process if it includes information on the servicing of the device (field-serviced by manufacturer-trained service persons or user-serviceable).

Insulation systems

The various global safety standards may have slightly different terms for the types of insulation required in a given piece of electrical equipment, but the insulation levels used in IEC-950 are common to many standards and are shown in Table 3 for reference.

Insulation Type	**Levels of protection provided**
Operational insulation *The insulation required for correct operation of the device*	0
Basic insulation *The insulation provideing the most basic protection against electric shock*	1
Supplementary insulation *Independent insulation applied in addition to basic insulation in case of a failure of the basic insulation*	1
Double insulation *Basic plus supplementary insulation*	2
Reinforced insulation *A single insulation system equivalent to double insulation*	2
Protective earth *Not insulation, but counts as one level of protection under certain circumstances*	0 or 1

Where to start

In this section, we will describe a process of determining the appropriate safety information you must have in your possession so you can communicate effectively with the suppliers of your safety-critical components. Incidentally, transformers comprise just one of several components that are scrutinized by safety agencies during the approval submission process. Other components include, but are not limited to, connectors, enclosures, printed wiring boards, wiring harnesses, interlocks, optical isolators, fuses, circuit breakers, overvoltage devices and capacitors. Any device that bridges the gap between user-accessible ports and hazardous voltages is likely to be subject to the scrutiny of an agency's safety engineer. The goal is to eliminate the hazard posed by a fault. Most safety agencies and standards require that the user be protected from hazardous voltages under single fault conditions. A fault is defined as the failure of any component that would expose the user to hazardous voltages.

To determine the applicable level of insulation required for a particular piece of equipment, you must be aware of the level of hazard the insulation will be called upon to protect the user from, and thus the level of protection required. The underlying principle of IEC-950 electrical safety is that two levels of protection are required to protect the user from hazardous voltages. This will still allow adequate protection in the case of a single-fault condition. The table below describes the various classes of circuits, their respective hazard level and the resulting level of protection required. To start:

1. Determine the working voltages at the port or ports that connect to the device. In general, each port must be assigned to one of the following groups:

Circuit	Definition	Hazard level
Primary circuit	Circuit directly connected to supply mains or other equivalent source.	2
Secondary circuit	No direct connection to primary power or mains.	-
Hazardous voltage secondary circuit	Secondary circuit operating at a hazardous voltage.	2
ELV (Extra-Low Voltage) circuit	Secondary circuit operating below a hazardous voltage (42.4 peak, 60V d.c.), but does not meet SELV or limited currect circuit conditions.	1
TNV (Tele-communications network voltage) circuit	Secondary circuit carrying telecommunications signals under normal operating conditions.	1
SELV (Safety Extra-Low Voltage) circuit	Secondary circuit operating under a hazardous voltage (42.4 peak, 60V d.c.) under normal and single fault conditions.	0
Limited current circuit	Secondary circuit from which the current that can be drawn is not hazardous under normal and single fault conditions.	0

2. Determine the environment in which the equipment is intended to operate. For example, is the device to be used in an office environment, a home or residence, or must it operate outdoors in extreme climates?

3. Determine the environmental conditions to which the safety barrier devices will be exposed. For example, is the device or equipment completely sealed, or does it have a fan? Are the safety barrier components subject to contaminants which are internal to the equipment, such as metal shavings or carbon powder? (Copy machines are notoriously "dirty" due to the toner they use.)

The last two items describe the *pollution degree* of the environment in which the barrier devices must reside. IEC-950 has three pollution degrees, Pollution Degree I, Pollution Degree II and Pollution Degree III. Products in Pollution Degree I are typically those which are hermetically-sealed, making products in this category relatively rare and specialized. The bulk of telecom equipment falls into Pollution Degree II which may have vented covers or fans to promote air flow, but are otherwise free from environmental hazards in a typical office environment (except for the occasional coffee spill). Products falling under the Pollution Degree III category, such as toner-based copiers, have the highest safety requirements due to the potential for conducting contaminants in their internal workings.

Inform your transformer supplier of the working voltages, their respective circuit type and the equipment's pollution degree so an appropriate choice of insulation methods may be made. Midcom has taken the stance that most of our products are used in equipment belonging to the Pollution Degree II category, so we apply the rules of that category unless we are told otherwise.

A typical arrangement for equipment that includes a telecommunications line interface is shown in Figure 21.

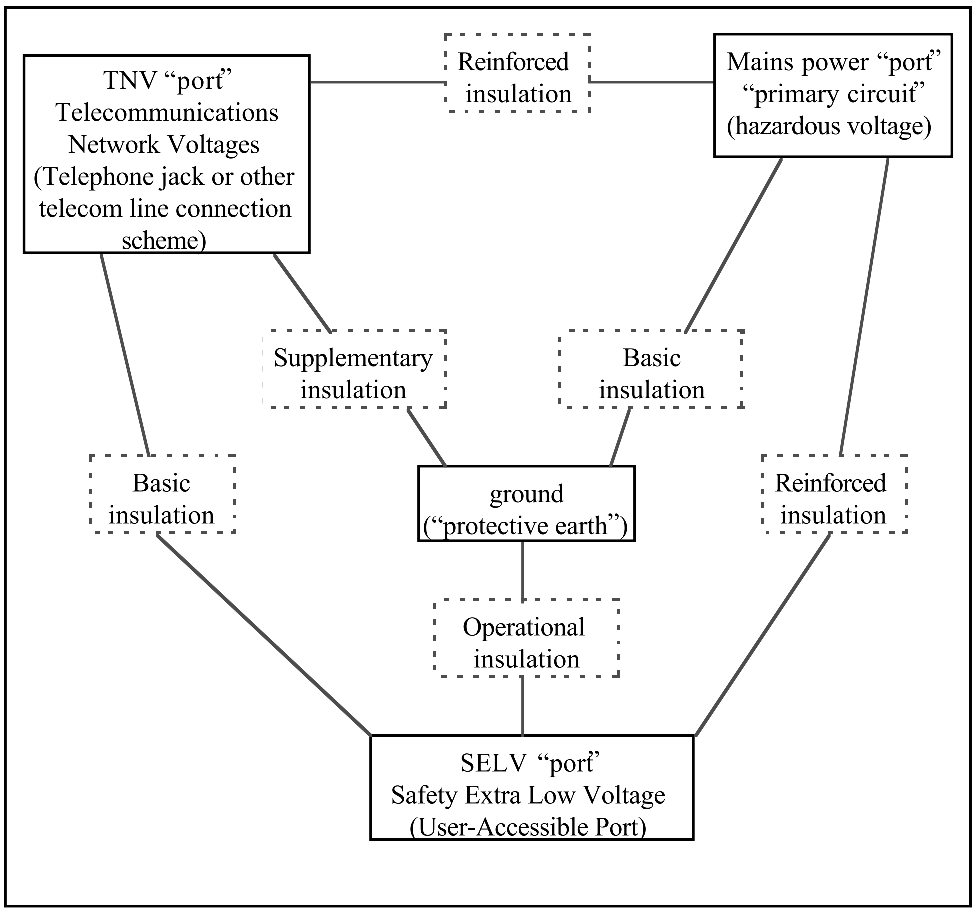


Figure 21

How working voltage, CTI and pollution degree influence transformer construction

In general, higher working voltages and higher (dirtier) pollution degrees require greater creepage and clearance distances. For example, a 120V working voltage under pollution degree II and CTI material group II requires 1.1mm creepage distance. The same pollution degree with a 250V working voltage requires 1.8 mm creepage distance.

Clearance distances and dielectric test voltages are determined by maximum DC, RMS, and peak working voltages. A high-voltage transient may arc from conductor to conductor along the clearance path. This means the clearance distance is also dependent upon the level of transient overvoltage the given insulation system will be subjected to, as well as the product's pollution degree.

In contrast to this, creepage distances are determined by steady-state DC or RMS working voltages (as opposed to transient voltages), but they are also dependent upon the product's pollution degree.

Note that simply meeting a dielectric test voltage is *not* a sole requirement for safety in the eyes of most regulatory agencies. Actual, measurable spacing distances are also required.

Non-safety regulatory requirements

Transformers have a significant role with compliance to non-safety regulatory requirements. In the telecommunications industry, transformers frequently form the first line of defense against problems associated with common-mode transmission line noise. Transformers provide an excellent means of making a balanced-to-unbalanced conversion and are well-suited to the task of preserving the balance of a transmission circuit. A transformer may also impact the return loss of a communications circuit. While this was discussed in an earlier chapter, we will expand on regulatory return loss requirements in this chapter.

Longitudinal balance

There are several ways of defining and measuring the degree of imbalance a transformer (or other device) may impart to a balanced circuit. One is to apply a common-mode signal to the DUT (Device Under Test) and measure the resulting differential voltage that results from the DUT's imbalance. This is known as the "L→M" or Longitudinal to Metallic conversion or method[1]. There also exists the "M→L" method where a differential signal is applied to the DUT and a resultant common-mode signal is measured at an appropriate point in the test circuit. Strictly speaking, the "M→L" method is not a measure of longitudinal balance, so a new term, *Transverse Balance* has been coined to describe this conversion.

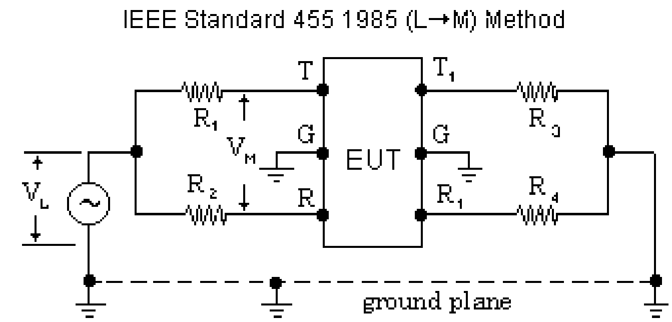
The "L→M" is more commonly-accepted throughout the world, and is the method described in ANSI/IEEE-455 which has been adopted by many countries.

Background on transverse balance

In the US, Africa and perhaps a few other countries, the "M→L" method still holds sway, but the "L→M" method is sometimes also specified. In observance of the proper terminology, FCC Part 68 rules have been recently revised, replacing the incorrect use of the term "longitudinal balance" with the new, correct term, "transverse balance" where it applies.

[1] The terms "longitudinal" and "metallic" come from telephone lore and probably pre-date the modern replacement terms "common-mode" and "differential."

The best way to see how these two balance tests are performed is to examine their test circuits. Figure 22 shows the ANSI/IEEE method of measuring Longitudinal Balance, while Figure 23 shows the FCC method of measuring Transverse Balance.



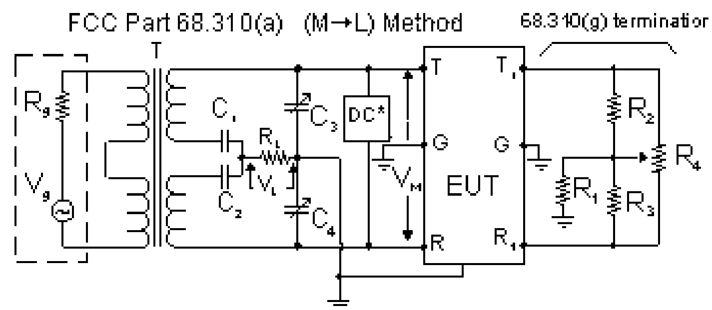
Longitudinal Balance = $20 \log_{10} \left| \frac{V_M}{V_L} \right|$ dB

R_1, R_2, R_3, R_4 = 368 ohm resistors precision matched for test set calibration within 0.01%

V_L = Longitudinal Voltage

V_M = Metallic Voltage (use of a balanced isolation transformer such as 671-0244 is recommended)

Figure 22



Transverse Balance = $20 \log_{10} \left| \frac{V_M}{V_L} \right|$ dB

C_1, C_2 = DC blocking capacitors
C_3, C_4 = trimmer capacitors
R_L = 500 ohms
T = Midcom 671-0323
V_L = longitudinal voltage

R_1 = 350 ohms
R_2, R_3 = 300 ohms
R_4 = 300k ohms

V_M = metallic voltage

Figure 23

Longitudinal conversion loss

To add to the confusion, another term has been adopted by the ITU (formerly CCITT) known as *longitudinal conversion loss* or LCL. LCL is essentially an "LÆM" method measured slightly differently, as shown in Figure 24. In most cases, the measurement is made with switch S in the closed position. Some equipment may require that switch S be open to make an appropriate measurement.

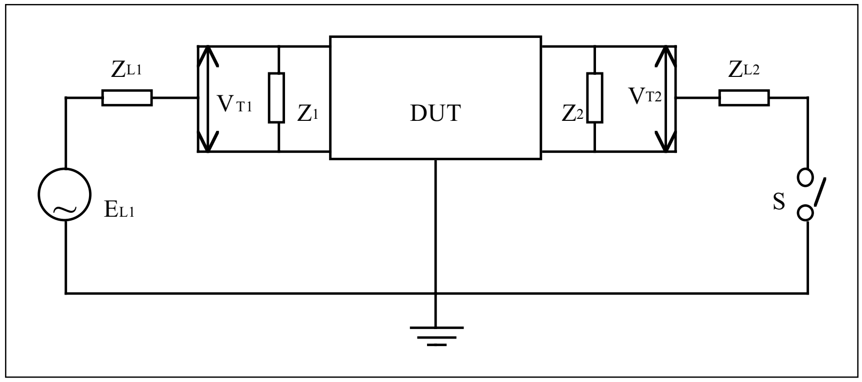


Figure 24

Return loss

Return loss may be specified as being measured with respect to a pure resistance or a complex impedance. Most telecommunications standards use either a pure resistance, a resistance in series with a capacitance, or a three-element complex network consisting of two series resistors, one of which has a capacitor in parallel with it. See Figure 25.

To prevent this document from quickly becoming obsolete, only a few of the more common return loss reference networks and their specifications will be presented. *These return loss specifications are listed only for the purposes of illustration of the variety of reference networks and limits.* Refer to the current relevant standard for the applicable return loss network and limits you need for compliance. You may also check for other Midcom technical notes that list return loss specifications by country and spec.

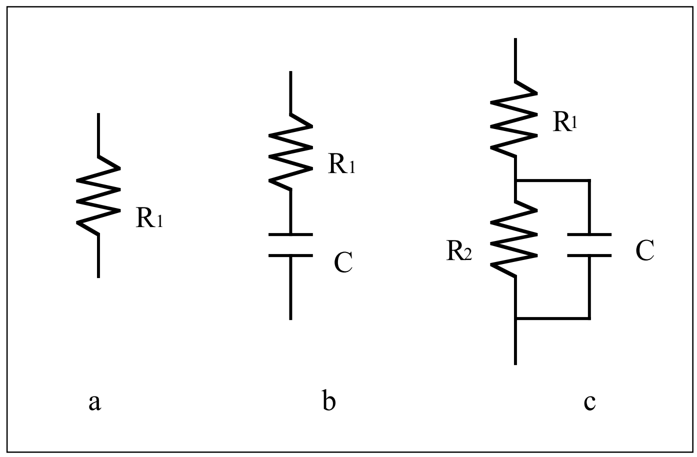


Figure 25

The following table gives examples of the range of return loss requirements for voiceband audio. Other specifications exist for frequency ranges other than voiceband audio.

Specification	Country	R1	R2	C	Limits	Frequency range
TBR-21 March 1996	Various European countries	270	750	150nF	6 dB 8 dB	200-300 Hz 300-4000 Hz
TS-002 1992	Australia	220	820	120nF	10 dB 15 dB	300-600 Hz 600-3400 Hz
TAS PSTN1 Issue 3 Rev 2, Feb. 1991	Singapore	600	--	--	15 dB 14 dB	300-3400 Hz

Table 5

Other requirements

Other requirements of a non-safety-critical nature will be documented here, as needed and upon request by readers.

Chapter 4 Construction

This section is devoted to the various core shapes used to construct transformers. Please keep in mind that these are just some of the shapes available. There are so many shapes used to construct magnetic components that we have selected only the more common ones used at Midcom. Most of these shapes are optimum for telecommunications and light power and several have been selected due to their ease of use in a high-volume manufacturing environment.

Toroidal cores

The toroid is the perfect magnetic shape from a performance standpoint. For this reason, it is a good shape to discuss first. Unfortunately, the toroid is one of the poorer shapes if one wishes to build products using high-volume automation.

The toroidal core is based on the geometric shape known as a torus. Strictly defined, a torus is an area swept out by a circular disk, as shown in Figure 26. By contrast, most toroidal cores are actually rectangular in their cross section, as seen in Figure 27.

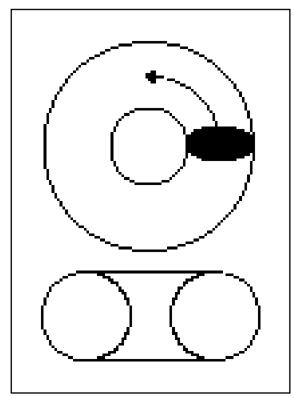


Figure 26

Toroidal cores can be wound such that wire completely covers their exposed area. Other core shapes either do not have this feature, or are impractical to wind this way. By making use of the complete core area we can allow two transformer windings to have the tightest coupling and thus promote efficiency. We can also spread the turns out along the entire core if we wish to reduce an inductor's distributed capacitance. These two advantages illustrate the toroidal core's key advantage over other core shapes. A third advantage is that solid, unbroken toroidal cores provide the highest permeability since they have no air gaps. This also helps to reduce crosstalk since higher effective permeability better contains the magnetic flux.

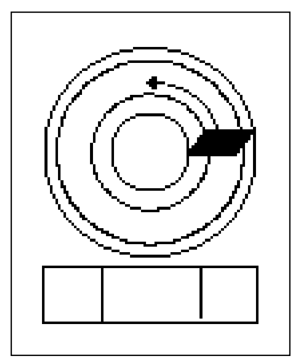


Figure 27

The downside of a toroidal core is that it is difficult to wind and mount. To place a winding on a toroidal core, you must "sew" the wire through the center hole. This means you must release the wire, push it through the hole, catch it as it comes out, and repeat the process. You can continue to do this until the center hole is full and you can no longer push wire though any remaining gaps. Completely filling the inside diameter is considered to be poor practice from a manufacturability viewpoint, but is fine for experimentation purposes.

There are machines that can automate the winding of toroids by applying wire in two stages: during the first stage, the wire is wound onto a temporary ring called a shuttle, then during the second stage the shuttle rotation direction is reversed, causing the wire to be wound onto the toroid. The shuttle ring has a cut through it, allowing it to be opened and placed through the center of a toroid. The shuttle also has a channel on its outer diameter to hold the wire. This method of winding, while ingenious, has the drawback that prevents the inside diameter of the toroid from being completely filled due to the space taken up by the shuttle ring. The shuttles can also hold only a limited amount of wire, so toroidal windings with many turns must be broken up into two or more applications. For more information, refer to US patent number 3601325, assigned to the Jovil Manufacturing company, the abstract of which may be viewed at

http://www.patents.ibm.com/details?patent_number=3601325

Figure 28 illustrates the winding of a toroidal inductor.

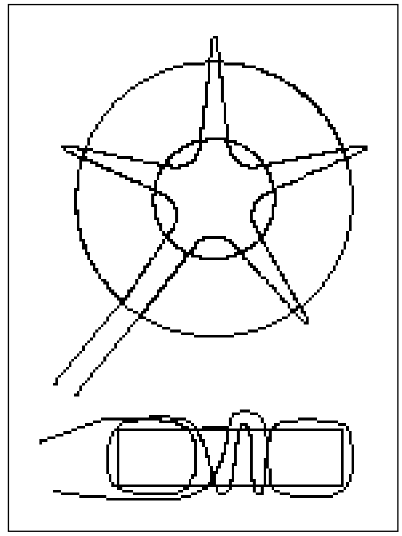


Figure 28

In addition to showing how the wire is routed through a toroid's center hole, the figure also illustrates how wires will eventually crowd the inside diameter, putting a limit on the size of the wire and the number of turns. When counting turns on a toroid, you can eliminate possible ambiguities about what constitutes a complete turn by counting turns on the inside of the core.

Another drawback to the toroid is that it may require a mounting board or header to make it mountable on a PC board. It is possible to make a toroid self-leaded such that the wire used to make the windings is left long enough that the part can be inserted into a PC board. Even in these cases, some sort of support or "header" may be required to hold the finished component to the PC board.

The E-based shapes

A more manufacturable shape of core, and probably the most popular, is the "E" shape. E cores may be paired with "I" pieces, other Es or modified to become F shapes. The E core is topologically equivalent to two toroids mashed together to form a single core. The windings are typically applied only to the center leg of the E shape, but may be placed on the outer legs for special purposes. Figure 29 shows how two toroids are equivalent to an E core in the magnetic sense.

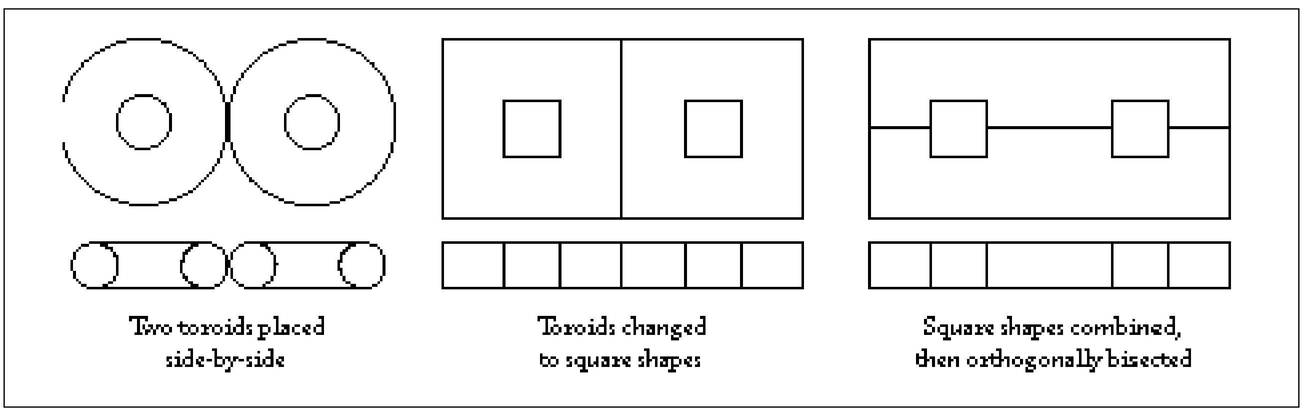


Figure 29

Note that the combining process yields a center-leg dimension (w_2) which is exactly twice that of the outer leg (w1). This is an important characteristic of E cores which allows equal flux distribution throughout the core as shown in Figure 30.

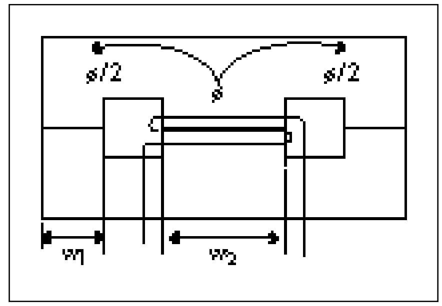


Figure 30

Although the most popular configuration for E cores results in w_2 being twice w1, there are a few special applications where this rule is violated. Ferroresonant transformers are one such example.

While it is possible to place coils on the outer legs of an E core device, it is generally impractical to do so except where the extra coil cost become a secondary factor to performance. If coils are used on the outer legs, they must each have half the turns of an equivalent coil placed on the center leg due to the outer leg width being half that of the center leg.

Other shapes based on the E core

A special lamination configuration known as the F shape was developed for the purposes of allowing its center-leg gap to be adjustable. As you'd expect, the shape looks like the letter F. See Figure 31.

Laminations of the F shape tend to be expensive because they don't make 100 percent use of the material from which they are stamped. Some shapes, such as EE and EI laminations, can be created such that no material is wasted during the stamping process. A die that produces such shapes is called "scrapless." When lamination sizes become very small, such those that result in a transformer less than about a half inch on a side, the benefit of scrapless laminations starts to dwindle. Ferrite cores are molded and thus their scrap is independent of shape.

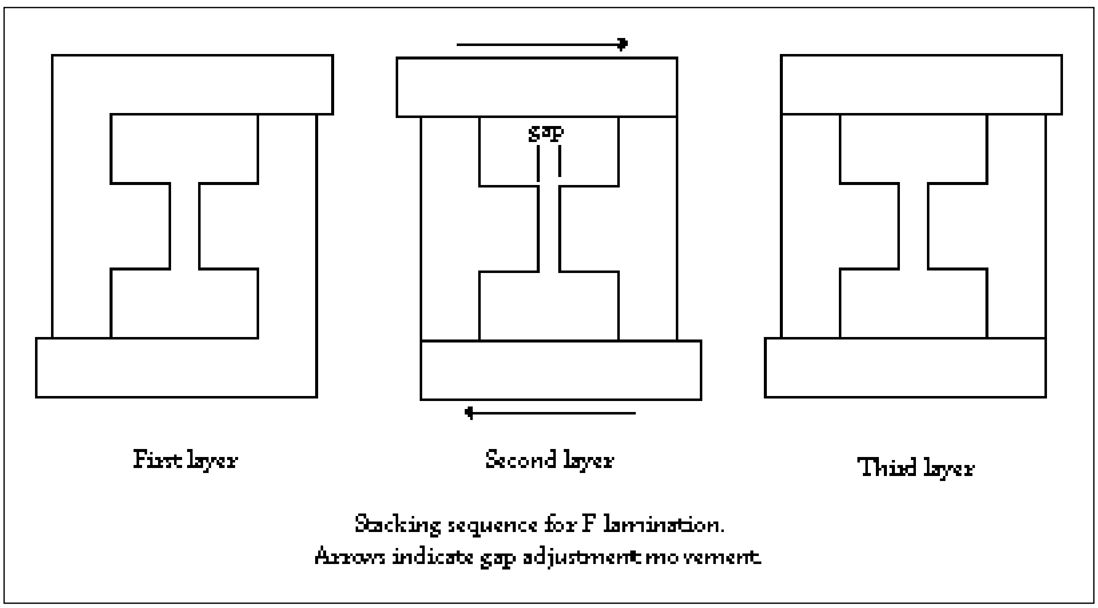


Figure 31

The I shape may be paired with the E shape, creating an E-I combination as shown in Figure 32.

Pilot holes are used by the lamination manufacturers to index and move the laminations during the stamping process. Some of the larger lamination sizes, such as those larger than two inches on a side, may have holes in their corners for the purposes of bolting the core stack together.

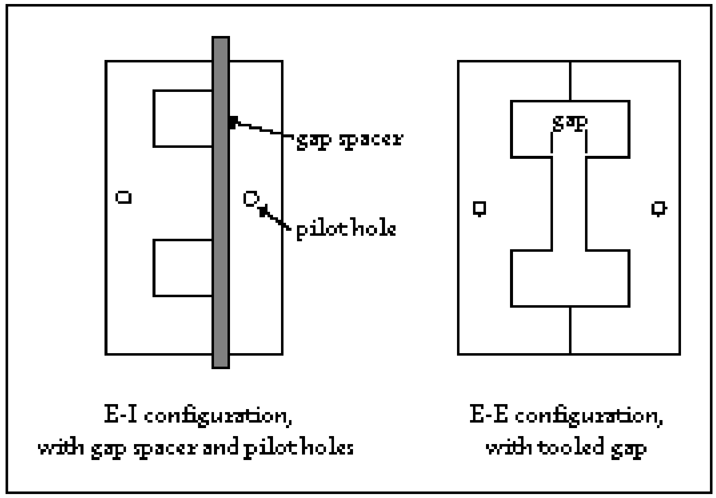


Figure 32

The U and C shapes

Due to an unfortunate circumstance in the English language, we have a shape that can be confusing when described verbally. The shape is the "U," but when used with another "U", it becomes the U-U or "double U" configuration. The difficulty comes when someone misunderstands "double-U" as being "W." Thankfully, there is no W core shape (as far as I know), so buyers need not worry that a shipment of W laminations will show up on the receiving dock as the result of somebody placing an order for double-U laminations. The C shape is the same as the U, but the C name is usually associated with tape-wound cores.

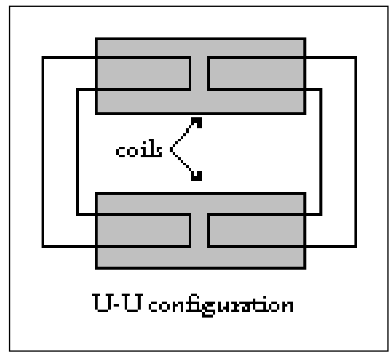


Figure 33

The U shape can approach the performance of the toroid because it can have more of its area covered by the coil. This is also a drawback because two coils are required in the U-U configuration. Figure 33 shows the U-U configuration.

The U-U configuration can also cause confusion with respect to polarity of the connected windings. I always have to think this through very carefully. Supposing coil A and coil B are identical coils: there are two ways of arranging them on the U-U cores.

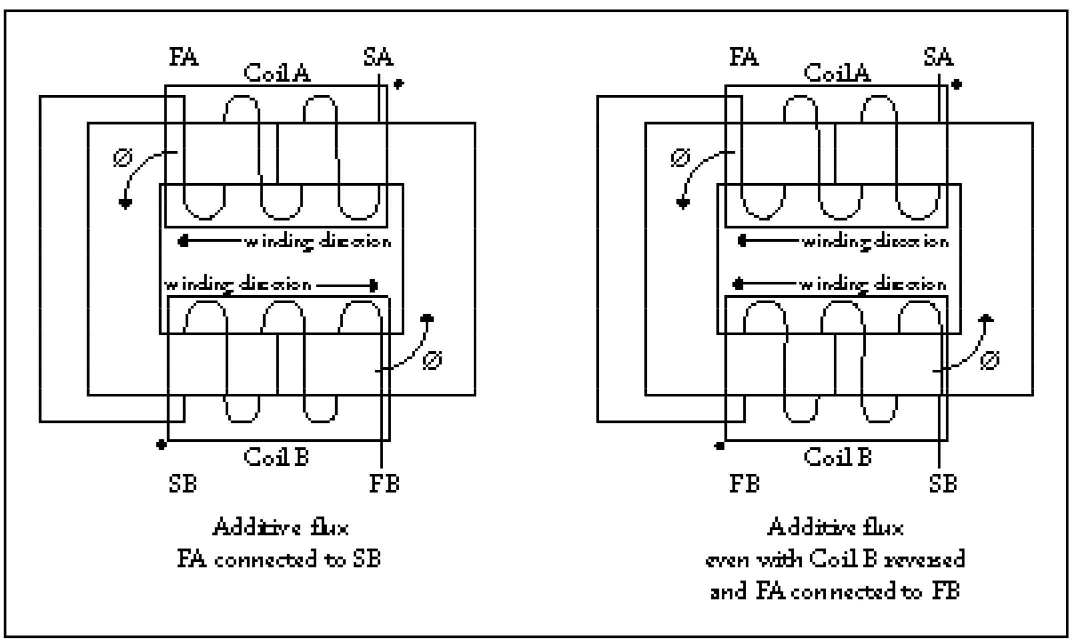


Figure 34

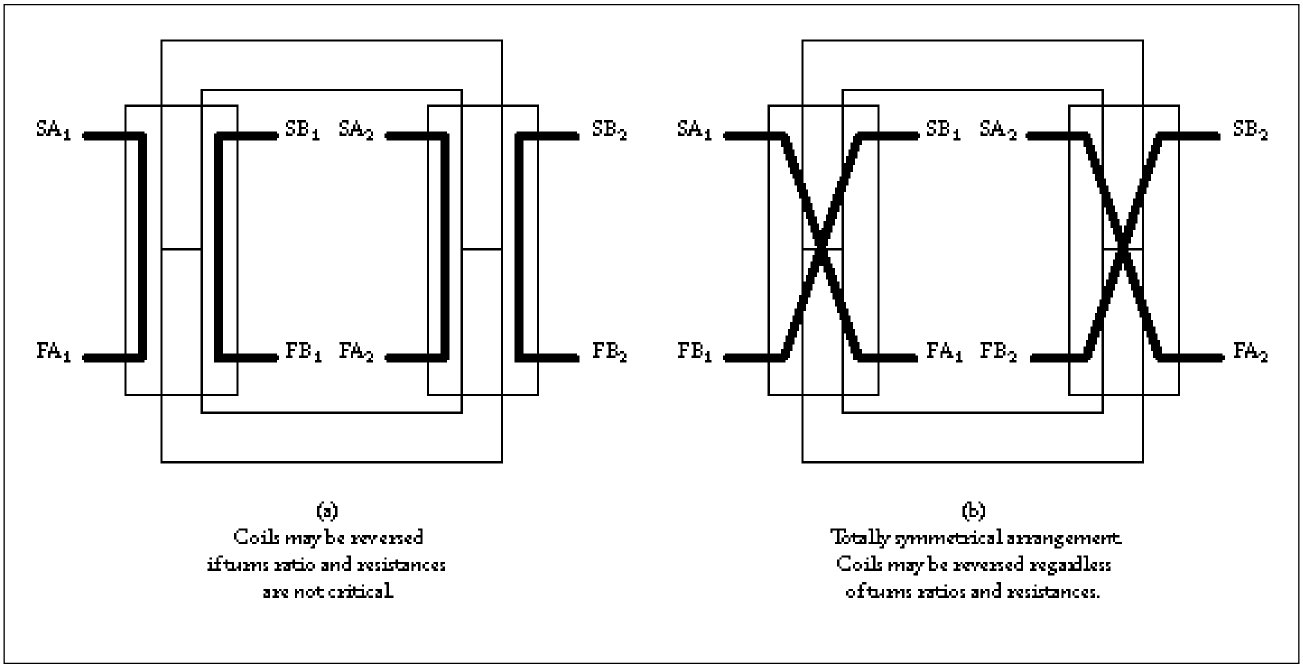


Figure 35

By reversing Coil B we have counteracted the effect of connecting the finish of the A coil to the finish of the B coil. Thus we have caused a physical finish to appear electrically as a start. We did this by reversing the sense of the magnetic field in coil B. In other core shapes where only one core leg contains a coil, there is much less opportunity for a coil to be incorrectly connected. The U-U shape is different in this respect and care must be taken when assembling transformers using this shape to make certain that correct polarity is maintained.

One way of preventing accidental polarity reversal is to wind the coils in a symmetrical fashion so the reversal of a coil makes no electrical difference. This can be easily accomplished if the transformer has an overall 1:1 turns ratio and if the coils are wound as shown in Figure 35(a). This technique will cause transposition of the A and B windings if one or both coils are reversed. If winding resistance or other differences between the A and B windings would cause the transformer to fail in the circuit, this technique cannot be used.

Another method—illustrated in Figure 35(b)—will work regardless of resistance or turns ratio differences, but is less convenient to wind from a coil manufacturing viewpoint.

Close inspection reveals that each coil contains a portion of the primary and the secondary. This is done to provide "tight" coupling between primary and secondary, reducing leakage inductance. If the primary coil were wound on one leg and the secondary on the other, we would find that the two windings would couple very poorly. Poor coupling may be acceptable for 50 or 60 Hz transformers, but even in those cases, the windings are split between the two legs with primary-to-secondary spacing and isolation maintained by a plastic flange in the bobbin. Having portions of both windings on the same coil also allows the coils to be identical. This reduces manufacturing costs since the economy of scale can be applied to volume manufacturing of a single coil instead of two.

Conductors

Conductors other than wire may be used to construct a magnetic component. Litz wire, copper foil and wires made of various alloys are some of the more common alternative conductors in the transformer industry.

One of the driving forces behind the development of different conductor types is the skin effect. The skin effect causes the effective conductor cross-section to be reduced as the frequency is raised. The skin effect causes the current density in the conductor to be redistributed due to the conductor's magnetic flux. This effectively squeezes the current to the outer portion of the conductor. The skin effect can be significant even at 50 or 60Hz which is why high-power lines are DC instead of AC. For transformers you could physically lift (or at least the transformers I could lift), the skin effect is more likely to show up as the frequency exceeds a hundred kilohertz or so. It is convenient to think of the skin effect as creating a skin depth or effective cross-sectional area which, although purely imaginary, is useful to illustrate how only the outer layer of conductor carries current at high frequencies. In actual practice, the current is distributed in a gradient so there is always some current at the center, but it becomes vanishingly small as the frequency goes up.

Litz wire

Skin effect may be countered through the use of Litzendraht or litz wire. Litz wire is a bundle of insulated wires, each of which is transposed in the bundle such that along a significant length, any single conductor will be, for some portion of its length, located in the center, the middle and the outer portion of the bundle. The transposition prevents any one conductor from being subject to the full forces of magnetic flux, which reduces the effective resistance of the entire bundle. The wires in a litz bundle must be insulated from one another for this to work. Were it not for this requirement, standard stranded wire could be used. Since the resistance of wire is a significant source of power loss, the term used to describe the resistance at elevated frequencies is Rac which is related to its zero-hertz equivalent, Rdc. In fact, you could consider Rdc to simply be a special case of Rac at zero hertz.

Litz wire bundles of 50, 100 or even more conductors are readily available. They are constructed by winding smaller bundles of six conductors into larger bundles. Those bundles may be "litzed" with other bundles to create progressively larger cables. As you might expect, litz wire can become expensive when large numbers of conductors are involved.

For more information on the benefits of litz wire, contact Kerrigan & Lewis (no web page as of this writing, but one is on the way) or New England Electric Wire Corporation at

http://www.neewcweb.com

Their literature contains the formulae you can use to calculate the effective resistances, both DC and AC, of various mixes of litz wire.

Copper foil windings

Copper foil is commonly used in switchmode power as a means of providing large current-carrying capacity in a small volume. Foil is harder to apply to a coil, particularly if high-volume automation is desired.

Foil must also be insulated to prevent turn-to-turn shorts. This is commonly done by sandwiching the foil with tape having enough width to guarantee complete coverage of the foil. The thickness of the copper foil may be chosen to handle the appropriate current, but thicknesses more than 0.030 inches (0.76mm) tend to be difficult to form to bobbins having a tight winding radius.

Foil windings are usually prefabricated in strips having an appropriate length to provide the required turns. Leadwires are attached to the strips during the prefabrication process and are dressed in place during the winding process. The leadwires must be chosen such that their current rating will not be exceeded, although the actual power dissipated in the leadwire is generally so small that undersizing the leadwire may not pose a problem.

The difficulty of dealing with leaded foil strips is one of the reasons that foil winding is not easily accomplished using automated means.

Alloy wire

Alloy wire may also be used to wind a transformer, but the alloys will have higher resistance per foot than will and equivalent gauge and length of copper wire. (Silver is the only metal with resistivity lower than copper) Certain alloys may have much lower temperature coefficient of resistance than copper, which is +0.4% per degree Celsius. Requirements that call for very little resistance change over temperature may be met by alloy wire, if the additional resistance can be tolerated. Alloy wire may be used to limit current or to withstand harsh environments, but will cost more than copper. Save for a very few special applications, alloy wire holds little advantage over that of copper wire.

Stick winding

If the magnetic core is not easily able to hold the wire on its own, a coil may be wound on a separate assembly which is placed onto the core. One of the oldest methods of accomplishing this is to wind the wire in sections onto a cardboard tube, then saw the tube into individual coils. Care must be taken to prevent the saw from wandering into the wire sections. The start ends of the coil are commonly buried into the winding and must be carefully picked out and hand-terminated to a more rugged leadwire. This method of creating a coil is called "stick" winding because the unsawed cardboard tube resembles a stick. Insulation is applied in sheets between windings, and sometimes during a winding application if that winding will have a high voltage present or must have low self-capacitance.

Bobbins (coil formers)

Bobbins, sometimes called coil formers, are usually made of plastic and may have terminals onto which the windings are terminated. The best possible bobbin would have infinitely thin walls with infinite dielectric voltage withstanding capability, be flame retardant, not cause tracking and cost nothing to produce. If such a bobbin had terminals, they would be perfectly solderable with no ability to move (thus guaranteeing coplanarity if the bobbin were surface-mount) and always be located where automated winding equipment could easily terminate on them. Clearly no real bobbin in existence can possess all these traits, so trade-offs must be made in each individual case based on the bobbin's application.

Even though there is no *perfect* bobbin, molding technology has advanced in many ways, even during my relatively short time in this business. We now have bobbins made with flame-retardant materials having excellent CTI ratings which take up less winding volume than ever before. Wall thicknesses down to 0.015 inches (0.38mm) allow very small bobbins to provide adequate winding volume. Surface mount bobbins, which caused considerable manufacturing difficulties when they were first introduced, are becoming easier to use from both a magnetics manufacturer's viewpoint and also from that of the board assembler.

Automation

Automation is a two-edged sword. It can reduce labor content as a function of capital expended. At the same time, it reduces flexibility as machines and materials become more specialized. A gentleman from the robotics industry once pointed out to me that human beings are the most flexible form of automation available. While this is true, it is my opinion that flexibility comes with a price of reduced consistency—possibly lower throughput. An operation or process dependent on the human element causes it to be only reliable within two or perhaps three sigmas on a bell curve. To reach five or six sigmas, (failure rates of a few parts per million) one must turn the processes over to automated machinery. I don't view this situation as a statement of the frailties of human nature as much as the appropriateness of capability. If I'm making only a few products per year, whether they are transformers, bicycles or TV dinners, a small support group can readily handle the variances due to human-built inconsistencies. If instead I were making several million products per year, I would not want my support costs to increase proportionally. I think this is the principal reason fast-food restaurants do not have table service. They simply can't serve a large volume of people and have the support costs associated with table service unless they could charge $10 for a hamburger (which, not coincidentally, is the price you pay at a restaurant with table service).

Automation provides greater consistency of product, but to keep automation development costs low, the components used to construct the finished product may be significantly different from those used to build non-automated product. Bobbins tooled for automated winding equipment may have special slots, knobs, and tie-off points that assist in the routing and termination of wire. Laminations may have gaps tooled into their center legs that take the place of manually-inserted gap spacers. Laminations may also be provided in pre-stacked and bonded sets in a process called staking, allowing them to be butt-stacked much like ferrite cores.

The benefit of automation is almost always determined as a result of labor saved, but occasionally the consistency improvements justify the costs incurred. The economic justification process is very straight forward. For example, $0.05 saved per unit may be applied to the cost of a machine or tool modification costing $10,000 making the break-even quantity equal to $10,000/($0.05 per-unit) = 200,000 units.

Effects due to temperature change

As discussed earlier, temperature will cause the resistance of a coil to change as a function of its resistance coefficient. Copper and aluminum, the most common metals used in transformer manufacture, have a relatively linear temperature coefficient of about +0.4% change in resistance per degree Celsius. Most core materials have a distinctly nonlinear shift in permeability with temperature, so it is important to understand how they behave over operational temperature ranges.

Permeability versus temperature: ferrite

The permeability versus temperature of high-permeability ferrite looks something like the curve shown in Figure 36.

Most magnetic materials have base permeabilities that vary with temperature in a fashion similar to that shown in the figure. Temperatures above 75° C were deliberately left blank because the curie temperature, the point above which magnetic materials lose their permeability, varies depending on the material. Most ferrites have curie temperatures around 125°C to 150°C. Some specialty materials have curie temperatures as low as room temperatures, or just a bit higher, making them useful in applications such as heat-sensing sprinkler systems.

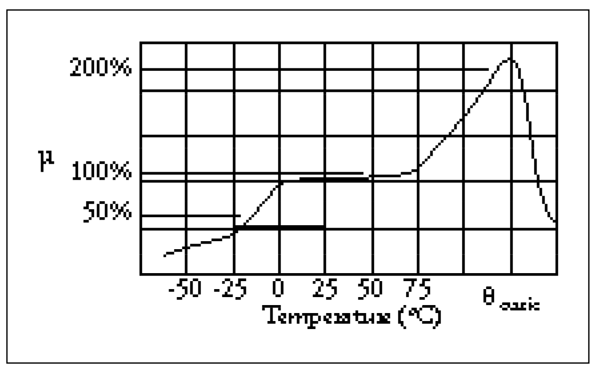


Figure 36

Remembering that inductance is directly related to permeability, we can see how circuit performance can depend on temperature swing. In the case of signal transformers, we may be able to overcome the problems of lost inductance at low temperature by simply choosing materials with high permeability. If we are already using the highest permeability possible or do not wish to add cost by increasing perm, we can add sufficient turns to provide the required inductance at the lowest relevant temperature. A common temperature range for telecommunications equipment is -40°C to +85°C. Based on the curve shown above, it is possible for a transformer or inductor to have an inductance shift of -50% to +150% of its room-temperature value. The most common way of reducing the overall inductance swing over temperature is to gap the core structure. Gapping has the effect of slicing or compressing the permeability excursion by a factor related to the ratio of the gap to the core's path length.

Permeability versus temperature: laminated sheet steels

For comparison purposes, Figure 37 is a chart showing how laminated sheet steels have a more linear relationship between temperature and permeability. Unlike ferrites, laminated sheet steel can have a negative permeability shift with temperature. Special thermally-stable steels can be manufactured that have relatively flat permeability shift with temperature. The range of these materials permeability can be seen between the solid and close-dotted lines below. Typical permeability shift for standard material can be seen between the dashed and solid lines.

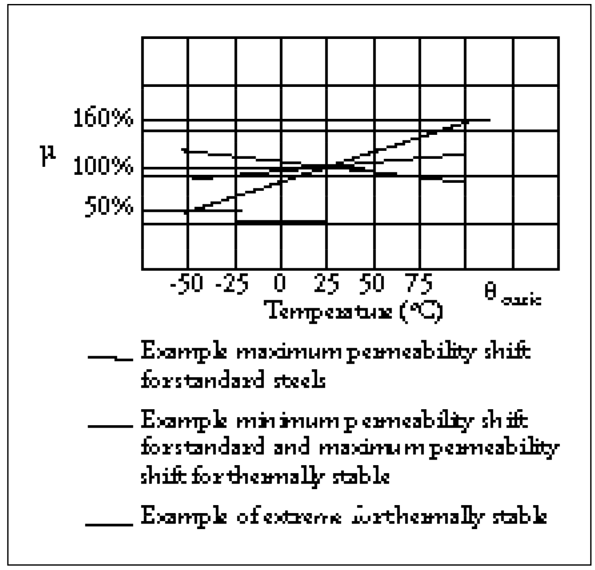


Figure 37

Spotting defects: what every component engineer should know

Dielectric failures

Probably one of the most important functions a transformer can provide is that of dielectric isolation. The isolation, also called hi-pot test (for high-potential) is usually tested by applying a voltage in excess of 500V to the device in question while monitoring leakage current for “rapid and uncontrolled increases.” Interestingly, many safety regulations do not specify the actual value of leakage current, although a few do. FCC Part 68 specifies an absurdly high value of 10mA. Most transformers are capable of containing leakage current to less than 500 microamperes, although a few larger devices may exceed this value due to capacitive coupling.

Many safety agencies specify a 60-second duration for the test. This is not economically practical for low-cost transformers, so many manufacturers have taken a bit of liberty with this by applying a higher voltage for a shorter period—say 20 percent more voltage for a shorter duration of perhaps one second. Other standards are written with practicality in mind. For example BABT’s interpretation of IEC-950 allows a one-second test at the rated voltage for high-volume production, although a one-minute test is performed when an article of the product is submitted for qualification testing.

The best way to spot an impending dielectric failure is to use a tester with an arc detector. An arc detector, such as the one found on Slaughter’s model 103 or 105, provides an audible indication of dielectric leakage. The arc detector allows monitoring of the leakage current and crackles as current begins to flow. In a typical scenario, the rapid, uncontrolled increase in current is preceded by an increase in crackling heard via the arc detector. Products that exhibit the increase in leakage should be considered suspect until otherwise validated.

Some dielectric testers include neon bulbs or other means of monitoring leakage current. It is important to understand that a small amount of leakage will always be present when testing the dielectric, and that transformers with large interwinding capacitances may exhibit a significant amount of leakage. If interwinding capacitance is so large that an AC test causes false indications of failure, an equivalent DC test may be applied instead. The most common way of making the two equivalent is to assume that the AC test waveform is sinusoidal and to apply a DC voltage equal to 1.414 times the RMS equivalent of the AC voltage. Some safety standards allow for the exchange of a DC test for an AC test while others do not.

Dielectric is a destructive test, similar in concept to drop-testing. One principle of reliability engineering states that a destructive test will cause a device failure, given enough test cycles. Thus it is possible that a transformer may pass dielectric testing before leaving the factory, yet fail upon reaching its destination. The key factor to note is whether the percentage of failed units in a population is increasing or decreasing upon subsequent retests. If the failure percentage increases from one test to the next, the design is clearly not robust enough to support a dielectric test at that voltage level. If the percentage of failures decreases from one test to the next, the remaining units in the population are all the more robust through the process of weeding out the "weak sisters." One can continue this process until the remaining population is as robust as desired, although one should be mindful that retesting has an economic impact on unit cost, both in terms of yield but also in time spent testing.

Open coils

Open coils are another common problem plaguing transformers and other magnetic components. Coils wound with fine gauges of wire are easily broken if mishandled and may warrant special attention during assembly. The process of soldering a coil doesn't do the wire any favors, either. Excessive solder dip time or too high a soldering temperature may cause the magnet wire to become brittle. Another symptom of extended dip time or excessive heat is crazing, the phenomenon that causes wire insulation to reflow and thus redistribute itself in ways that can leave thin spots where insulation is insufficient to protect the wire.

Open coils are easily found by a simple continuity test. Other tests, such as inductance and resistance, are also ways of culling open coils from a production lot. Wires may be nicked or cut during core assembly; coils may open at stress points if they are dropped or subjected to similar forces—especially if the forces are applied to the point where the wire is terminated to the terminal. Handling during additional assembly steps, testing and packaging are all potentially detrimental to the continuity of a coil.

Assuming the transformer manufacturer is astute enough to avoid these pitfalls, there are other opportunities to generate failures when the transformers reach their destination. Careless unpacking and poor handling practice are two common sources of broken coils. While transformers can be made quite robust, this does not come without a price, so unless you are willing to pay for bullet-proof products, it pays to follow a few simple guidelines. These may be obtained from the Midcom web site at

http://www.midcom-inc.com/technology/technotes/tn67.pdf

but a partial list is reproduced here. These guidelines may not apply to all magnetic components, but you should heed their words of caution and try to avoid violating them unless you are certain that they do not apply.

Tips to prevent magnetic component breakage

- Store them in their original shipping containers.
- Don't place them in bins or otherwise allow them to stack on top of one another.
- When hand-placing them, do not hold more than one in hand at a time. (This prevents the sharp terminals of one transformer from damaging the windings of another when pressed together in the palm.)
- Do not place or otherwise store them in plastic bags.
- If you need to return magnetic components to the manufacturer, try to reuse the same packaging material in which you received them. If you can't reuse the material, try to keep the units from bouncing into one another by securing them with styrofoam or similar material. Anti-static packaging is generally not required since transformers are quite robust with respect to ESD concerns, but consult the transformer supplier before assuming that units can be shipped in static-prone containers.

Inductance failures

As a rule, inductance failures aren't all that common, but when they do occur, they are usually due to another more serious problem. Inductance failures may be due to cracked cores, bent laminations, shorted windings, potting material that wasn't completely cured upon shipment or out-of-spec core material. To be meaningful, inductance must be tested at an appropriate frequency and drive level and in the correct equivalent (series or parallel). Some inductance bridges, particularly the older ones, don't provide a steady output level, so don't be shy about measuring the bridge's drive level at its test points by using a floating (battery-powered) voltmeter. You may be able to use an unbalanced (one lead grounded) voltmeter, but note that the bridge may have a "high" and "low" terminal and connect your voltmeter accordingly (high to high, low to low). The easiest way to determine if you can use an unbalanced voltmeter is connect the meter to the bridges test points, note the reading, then reverse the meter connections and note the new reading. If the two readings are substantially different, or the bridge complains that it can't supply adequate voltage, you're probably using an unbalanced voltmeter. You could assume that the higher reading is correct, but check this assumption with a battery-powered meter—just to play it safe.

If the inductance level is specified with DC current applied, you will obtain a different value, usually higher, when you test without DC present. Some transformers and inductors are specified to meet a particular inductance range from zero to a maximum value of DC. Unless the magnetic device contains a nonlinear component (such as a diode) or a permanent magnetic (unlikely), the direction of DC is unimportant. Use caution when testing inductors with DC. Current does not change instantaneously through an inductor, and it will create high voltages if interrupted as the magnetic energy is converted into electrical energy. Try not to be in the path of such current as even smallish inductors can contain significant energy. The energy in the inductor is proportional to LI^2 so a doubling of the current unleashes a fourfold increase in energy.

The inductance of well-designed magnetic components shouldn't shift downward by more than about 15 percent from their no-DC value to their full-rated DC inductance. If you see more than 15 percent shift, but the inductance is well within spec, you probably don't need to be concerned. If, however, the inductance shifts 30 percent or more and is only a few percent over the minimum value at full-rated DC, you should be concerned that there are failures within the lot as the design may be suspect.

Shorted turns

Shorted turns are surprisingly difficult to detect under certain circumstances. If a transformer or inductor contains a large number of turns of fine wire and only one turn is shorted, the result may be only barely noticeable as a slight decrease in the Q of the inductance. Here's why:

Suppose a transformer having 1000 primary turns and 1000 secondary turns has a shorted turn on the secondary with a resistance of 0.1 ohm. The shorted turn acts as a third winding with a 0.1 ohm load. That single shorted turn reflects back onto the other windings a resistance of $0.1\ (1000)^2$ or 100K ohms. Depending on the losses in the transformer core, the additional 100K may provide no measurable difference. Figure 38 illustrates this example.

The above condition is difficult to spot, even for an experienced magnetics designer. One method of determining shorted turns is to apply a pulse the magnetic device, then measure how long it "rings." The longer the ringing during, the less likely there is to be a shorted turn. To use this method, one must have a "known good" inductor or transformer to serve as a comparison.

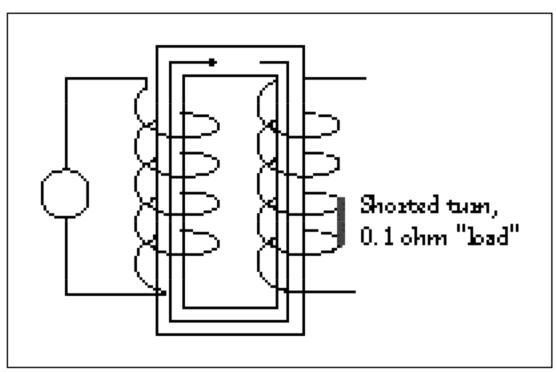


Figure 38

If the turns count is small, coupling is adequate and the number of shorted turns represents a significant number of the total turns, the transformer or inductor will behave significantly different from a similar non-shorted product, making the defect easy to detect.

Turns may become shorted if the wire build of the coil is too large causing the core to scrape through the wire insulation and thus bridge a turn or two. Hot solder may splash onto the coil and melt the insulation which can also short turns.

Conclusion

This concludes the chapter on construction practices, although more content will be added later as methods and materials change.

Chapter 5 Applications

In this chapter we will discuss the applications of transformers and inductors and how their characteristics play a role in system performance.

Analog telecom modem

Low-speed, no echo cancellation (V.22bis, V.29)

Low-speed modems—perhaps better described as modems that do not employ echo cancellation—place very minimal requirements upon the line coupling transformer. Signal distortion is generally not an issue unless THD exceeds a percent or two at 600 Hz. The most popular of these low-speed standards carry ITU designations of V.22bis and V.29. These data transmission standards describe 2400 bps full-duplex and 9600 bps half-duplex, respectively. Both standards use QAM signaling schemes that restrict the bulk of their energy to a frequency range of about 600 to 3000 Hz. Because the requirements are relatively loose, any transformer having 200mH or more with less than two percent THD and resistances less than about 150 ohms on each winding (assuming 1:1 turns ratio) will work well in a low-speed modem application. It is for this reason that many low-speed modem transformers can carry the DC loop current provided by the central office. A transformer required to carry loop current will perform more poorly than one of an equivalent size that need not carry current. In the case of low speed modems, the poorer performance is not significant and the line interface circuit can be made simpler through use of a DC-capable signal transformer. Many people assume that a DC-capable transformer will cost more than a "dry" or no-DC transformer, but in practice the reverse is true. Their assumption is based on the notion that a DC-capable transformer "does more" than one which can't. Since DC-capable cores may be butt-stacked their assembly time is less and thus they can be built for lower cost than "dry" transformers whose laminations must be interleaved one at a time.

Transformers for low-speed applications can be built using cores of low-grade silicon steel, although nickel steel may be used in critical applications where the modem signal level is lower than normal. Nickel steel performs better at low drive levels than does silicon steel. Leakage inductance is not critical as long as it is below 20 mH or so.

Medium-speed using echo cancellation (V.32, V.32bis)

The medium-speed technologies were the first to use echo cancellation and thus require reasonably faithful transmission of the modem signal. The two transmission standards that fall into this category are V.32 and V.32bis. V.32 is a 9600 bps scheme using trellis-coded modulation (TCM), that can drop back to 4800 bps QAM if required. V.32bis is sort of a combination of the V.17 half-duplex fax scheme combined with V.32 to create a 14,400 bps transmission standard using TCM. These speeds generally call for transformers to have no more than about –76dB THD at 600 Hz. This specification precludes many DC-capable transformers, except for ones larger than 1.5 inches (38mm) on a side and approximately 0.7 inches (18mm) tall. Almost all transformers for this application must be built with nickel-steel laminations or high-permeability ferrite core material.

Typical inductances for transformers meeting this application's requirements start at 0.5 henry minimum. DC resistances are typically less than 150 ohms on each winding to keep insertion loss to a reasonable figure. Leakage inductance is not terribly important as long as it is below about 20mH (to minimize high-end frequency response rolloff). Typical leakage inductances range between 5 and 15 mH.

High-speed using echo cancellation (V.34,V.90)

The high-speed transmission schemes begin to place significant constraints on transformer performance. The maximum allowable transformer distortion is around –82dB at 600 Hz for V.34 (33.6kbps). Varying numbers have been ascribed to the V.90 scheme, but at present we assume that –86dB THD is appropriate and frequency response must be down no more than –3dB at 10Hz to support V.90's Pulse Code Modulation (PCM) scheme. These requirements all but mandate that a dry transformer must be used and it must contain a core that is compatible with the low-distortion requirements. Typical core materials suitable for this would be 80 percent nickel-steel, low-loss/high-permeability ferrite or possibly amorphous materials. Inductance should be at least one or two henries to support V.34 and at least five henries for V.90 (for support of the –3dB, 10 Hz response requirement). DC resistances are typically less than 150 ohms for each winding (again, assuming a 1:1 turns ratio). These constraints begin to rear up as the allowable size drops into the PCMCIA-compatible realm. Leakage inductance isn't particularly critical as long as it is below 10-15 mH. More leakage inductance may be allowed, but less is generally better.

Digital telecom

DDS

DDS, or Digital Data Service (also known as Dataphone Digital Service or Digital Data System) is a digital service popular in North America that offers digital connections up to 56kbps full-duplex. DDS is very popular as a means of shipping data between financial institutions, as part of a private company's Wide Area Network (WAN) or as a means of providing full-time internet access. It is also used for telemetry and probably by ATMs (Automatic Teller Machines—*not* Asynchronous Transfer Mode). DDS signaling rates may range between 2400 bps and 56kbps (although implementations at 64 and 72 kbps also exist), but it is the lowest rate which determines the minimum transformer inductance. The frequency range of a variable-rate DDS signal is 100 Hz to 112 kHz, but in practice almost all connections are fixed-speed and most of those are 56kbps. If full-range rate support is desired, the minimum inductance must exceed 200mH. If only the 56 kbps or faster rates are to be used, the inductance can drop to 40mH. Leakage inductance for the 56 kbps rate ought to be less than about 45uH and DC resistance should be compatible with the DDS line impedance of 135 ohms.[1] Typical values range from 10 to no more than 25 ohms for each winding (again, assuming 1:1 turns ratio), depending on how much insertion loss can be tolerated. The frequency range precludes all but the lowest loss laminations, making high-permeability ferrite a good choice for core material. The frequency response is defined by a Bellcore standard and is ±1.0 dB from 100 Hz to 112 kHz. Despite ISDN's faster rate and dialup access, DDS is still a viable product area and will probably remain so for years to come.

[1] The impedance of a telephone line over the voiceband range is nominally assumed to be 600 ohms although this is based on the placement of two AWG 10 wires spaced on foot apart. Modern telephone wire is assumed to be 135 ohms for DDS signals. At frequencies above 100kHz the impedance is considered to be 100 ohms. Actual impedance varies widely but decreases with rising frequency due to the capacitive nature of the wire pair. See Midcom Tech Note #23 for empirical impedance measurements made over voiceband frequencies.

T1/E1/ISDN primary rate

T1 transformers

In Japan, Hong Kong and North America, we use T1 (sometimes denoted T-1), a transmission standard which wraps 24 voice channels sampled 8000 times per second with an eight-bit word "depth" into a high-rate multiplexed signal. Add on a few bits for overhead and framing and you get a total data rate of 1.544 Mbps. (A practice known as "bit stealing" provides supervisory control and monitoring of the connection to determine off-hook condition, ringing, busy, etc, so the net bandwidth available to the user is only 56 kbps) Most of the T1 energy is centered at 772kHz, but other requirements of the signal cause T1 transformer to have a frequency response range of 50kHz to about 2MHz. The minimum inductance for the line-side windings is around 0.6mH. The leakage inductance measured at the primary winding should be no more than a few microhenries. The line-side winding is usually center-tapped to allow phantom-feeding of DC current. The DC current is commonly used to power remote T1 repeaters which regenerate the signal every few kilometers. Since the DC is fed via a center tap, there is ideally no net DC flux in the transformer core. Mismatches caused by loop length differences or wire tolerance can introduce enough imbalance such that 1mA more current may flow in one half of the transformer when compared to the other half. This is magnetically equivalent to 0.5mADC flowing in the entire line side, so the line-side inductance should be capable of supporting 0.5mADC without a significant drop in inductance.

Interwinding capacitance may be important from a safety perspective. Typical acceptable values for interwinding capacitance are around 30pF. Resistances must be kept fairly low since the T1 line impedance is 100 ohms. Line-side resistances should not exceed five ohms and should typically be in the range of 0.5 to 2.0 ohms.

The turns ratio of the transmit side is commonly configured to provide a step-up to the T1 line. The T1 line code is a 6V peak AMI (Alternative Mark Inversion) scheme which means that every "1" bit has a polarity opposite that of the last "1" bit. The "0" bits are represented by a voltage of zero volts. Since the "1" bit voltage is 1V higher than 5V (or 1V lower than –5V), a step-up transformer is a convenient way of providing the correct voltage without requiring a special power supply. Common turns ratios for T1 transmit transformers are 1:2CT, 1:1.41 and 1:1CT. The equipment-side resistance should be scaled by the square of the turns ratio such that its value is about the same as the line-side resistance. For example, a transformer with a line-side resistance of 1.5 ohm and turns ratio of 1:2CT should have an equipment-side resistance of about $1.5/(2^2)$ or 0.375 ohms.

E1 transformers

The E1 transmission standard is much the same as T1 except that it carries 30 channels of 64 kbps data. This aggregate amount with a small amount of overhead adds up to 2.048 Mbps of total data throughput. There is very little difference between a T1 transformer and an E1 transformer and many times they can be used interchangeably. Some E1 installations have stringent return loss requirements and may need more inductance to support them. Some E1 circuits may use 75 ohms coaxial cable or 120 ohm twisted pair wiring. In these cases, the turns ratios must be adjusted to match the impedance of the line to the driving impedance of the chip-side circuitry.

ISDN Primary Rate

The ISDN primary rate is actually either T1 or E1 depending on the country in which it is being used. It sometimes carries with it additional return loss requirements. See the information on T1 and E1 above for details.

HDSL

HDSL stands for High-rate Digital Subscriber Line. HDSL has variable rates depending on need. HDSL is being used to replace T1 and E1 circuits because it does not require repeaters. Common rates for HDSL are 160, 288, 360, 400, 416, 528, 576, 784, 1168 and (eventually) 1544 and 2048 kbps. HDSL uses a line code known as 2B1Q (see information to follow on the ISDN "U" interface transformer regarding the 2B1Q line code). The data rates are determined by the HDSL IC manufacturers, although 784 and 1168 kbps are common to most. A few IC manufacturers allow the same transformer to be used at multiple rates, but most require different transformers for the different data rates. In general, the lower the rate, the higher must be the transformer's line-side inductance. For example, the 160 kbps rate requires a transformer with 8mH while the 576 kbps rate requires 2.5 mH. The ICs are developed to work with a specific inductance value which is usually required to land within ±5% or ±10% of the nominal value. The HDSL technology uses Digital Signal Processing (DSP) to cancel echoes and needs to know the inductance in order to compensate for the echoes on the line. Leakage inductance also follows the data rate, but in this case, higher rates require lower leakage inductances. Typical leakage inductance values for the 784 and 1168 kbps are 20 uH and 11 uH respectively.

Distortion also plays a significant role in HDSL performance, just as it does in analog modem performance. Here, too, the faster data rates require better distortion figures. A typical requirement is –70 dB THD max. at 40 kHz, +14 dBm drive level.

The HDSL transformer may be required to carry DC to a remote point. Commonly, as much as 160mADC may flow through the transformer. The line-side winding is usually split into halves so the battery-feed current can be extracted without unbalancing the line. Resistances are typically a small percentage of the line impedance. Values between 0.2 and 1.0 ohm are common.

ADSL/RADSL

ADSL stands for Asymmetric Digital Subscriber Line. A close relative of HDSL, ADSL provides more bandwidth in one direction than the other. This makes it a good candidate for the provision of video and internet services. Typical configurations call for 1.5-6 Mbps "downstream" toward the subscriber and an "upstream" channel of 16 to 640 kbps. Many ADSL implementations are required to work in the presence of standard voiceband telecom service called POTS or Plain Old Telephone Service. For this reason, some ADSL vendors have configured their systems to use filters to keep the ADSL and POTS signals separated. These filters are classic inductor/capacitor arrangements called ADSL/POTS splitters. The Central Office (CO) end of the line and the Customer Premises Equipment (CPE) end may employ transformers with differing characteristics. The CO side is sometimes called the ATU-C, or ADSL Transmission Unit - while the CPE side is sometimes called the ATU-R (the "R" presumably referring to "residence").

ADSL transformers have line-side inductances ranging from a few hundreds of microhenries to a few millihenries. They do not need to carry DC but are gapped anyway to control their inductance within a ±5% to ±10% range. Leakage inductances are roughly proportional to line-side inductances, ranging from a few microhenries to a few tens of microhenries.

ADSL systems may employ echo cancellation in the frequency range where the upstream and downstream signals overlap, making distortion a critical factor. Typical distortion requirements are –85 dB maximum THD for the CPE end and –80 dB THD for the CO end; both measured with a 15Vp-p signal at 100 kHz.

An rate-adaptive version of ADSL known as RADSL allows varying data rates depending on line condition or as a means of allowing the user to choose the rate that is appropriate for the desired level of service. For more information on ADSL, refer to

http://www.adsl.com

ISDN

The ISDN "U" interface

The ISDN "U" interface is the connection that carries the ISDN signal between the CO and the "User's" premises or CPE (at least that's how I remember it). That connection can carry DC to power the user's equipment in case of a power failure there. The signal bandwidth is approximately 2kHz to 200kHz over which range the transformer is usually flat within 0.5 dB. The line code is 2B1Q, (also used by HDSL) which means that two binary bits of data are converted into a single four-level signal before being transmitted. Figure 39 shows an example of input data and the corresponding voltages seen on the line.

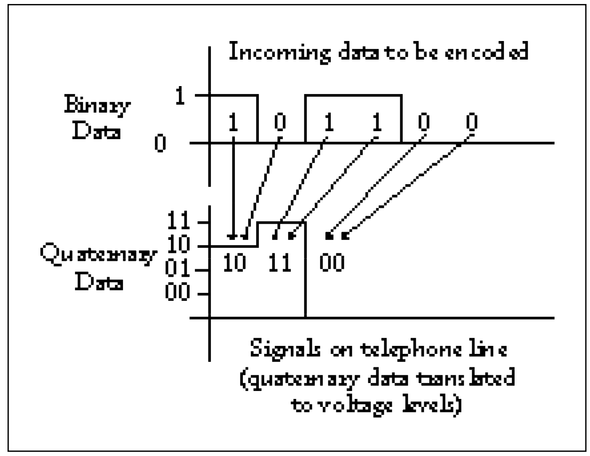


Figure 39

The ISDN "U" system relies on echo cancellation since all communications takes place over a single pair of wires. A total of 160 kbps full-duplex transmission on a single twisted-pair is provided, with user-accessible channels including 2 "B" channels, each at 64 kbps, one "D" channel at 16 kbps, and an additional 4 kbps for loop maintenance. 12 kbps of bandwidth is reserved for framing. Using the 2B1Q Line code, pairs of binary bits are coded into 1 of 4 quantum levels for transmission at 80k symbols/sec (hence 2 Binary/1 Quaternary). Reliance on echo cancellation requires low distortion. One typical spec is –53 dB maximum at 500 Hz with a drive level of 5Vp-p. Resistances on the line side are typically 5-20 ohms. Resistance of the secondary winding is typically scaled by the square of the turns ratio. Leakage inductances are typically 15-50 uH measured at the line-side winding. Turns ratios depend on the chip manufacturer's requirements, but many have chosen to step up the voltage to the line by a factor of 25 to 50%. The line-side winding is usually split into halves so the battery-feed current can be extracted without unbalancing the line.

The ISDN "S" interface

The "S" interface takes place over a four-wire pair sometimes employing phantom current feed to the remote devices. Since transmit and receive channels occupy their own wire pairs there is no need for echo cancellation, thus there is no distortion requirement. The specification for the "S" interface requires good coupling in that the ratio of magnetizing inductance to leakage inductance is quite high. The typical line-side inductance requirement is 22mH minimum. Typical leakage inductance is 8uH maximum. The turns ratio is dependent on the chip manufacturer but most have settled on either a 2:1 or a 2.5:1 equipment-to-line ratio. The line impedance for this interface is considered to be 100 ohms, making the driving impedances 400 or 625 ohms for the two popular turns ratios of 2:1 and 2.5:1 respectively. Distributed capacitance can be critical in ISDN "S" interface transformers due to a minimum impedance template which must be met when the transformer is tested with an artificial line connected to its line-side terminals. DC resistances must be appropriate for the line impedance and turns ratio. Typical values for the line-side resistance range from five to seven ohms with the equipment side resistances ranging from seven to 15 ohms.

Telecom/Voice

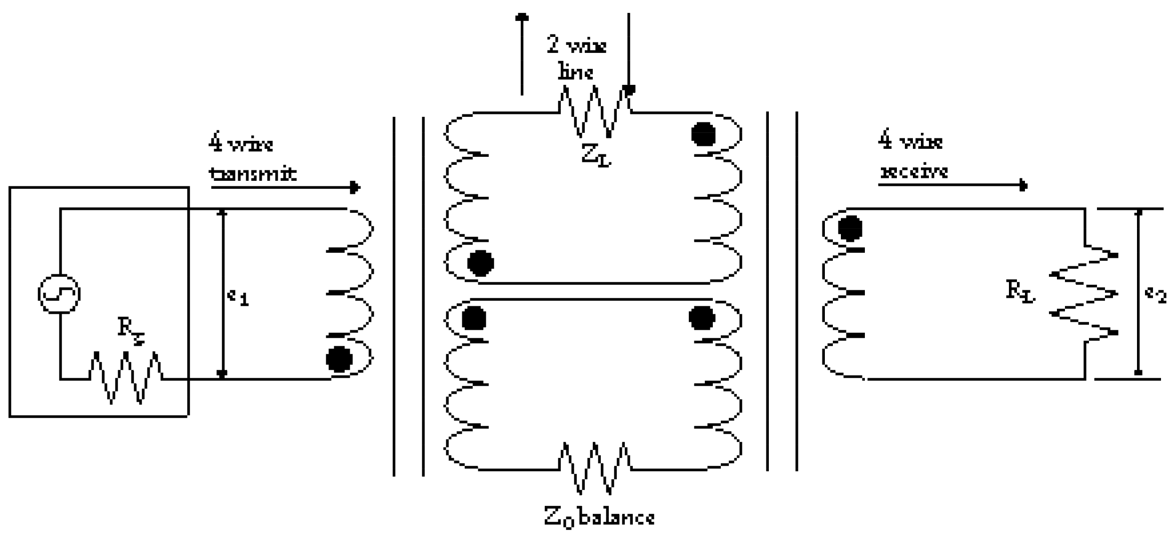
Standard couplers

There are many types of telephone line-coupling transformers. Bridging transformers are used to monitor the line and therefore must present a very light load to the line impedance. Such transformers usually have moderate to high inductances, typically in the range of five to 20 henries.

The relatively high inductance coupled with the high input impedance of the monitoring amplifier means the transformer can have high DC resistances without impacting performance. Typical resistances for bridging transformers can range from 50 ohms to a few hundred ohms. Frequency response is usually measured from 200 to 4000 Hz with a typical response requirement of ±0.5 dB. The turns ratio of a bridging transformer is usually 1:1 but may be adapted to match varying impedances if desired. Insertion losses are typically 0.5 to 2dB. Due to the similarities in construction, many high-speed modem transformers can be successfully adapted for use as bridging transformers.

Bridging transformers need not carry DC current. In fact, most bridging transformers would saturate in the presence of significant DC current. Bridging transformers are routinely decoupled from DC through the use of a suitable capacitor of, for example, 1uF. Larger capacitor values will provide better low-frequency response.

Line-interface transformers or line couplers are designed to work with either “wet” (DC-carrying) or “dry” (no-DC) lines. If the line is wet, the transformer can be designed to carry the DC which makes for a simple line interface design. Wet transformers will have much less inductance due to the air gap in their core which is required to prevent DC from causing core saturation. Typical inductance values for wet couplers required to carry 30mADC or more ranges from 150mH to perhaps one or two henries, although the large inductance values necessitates a large transformer of perhaps one inch (25.4mm) or larger on each side. Resistance is typically 20 ohms to 120 ohms on the primary (line-side) winding. The secondary side resistance is usually scaled by the square of the turns ratio, but since many line couplers have 1:1 turns ratios (or values close to 1:1) the secondary resistance is usually about the same value as the primary resistance. Frequency response is usually subject to roll-off at the low end due to limited primary inductance. Typical values of –3dB or –4dB



Trans-hybrid loss (THL) = $20 \log_{10} \left| \frac{e_1}{e_2} \right|$ dB

R_S = Four-wire source impedance

R_L = Four-wire load impedance

Z_0 = Two-wire line impedance

Z_L = Balance network impedance

e_1 = Four-wire transmit voltage

e_2 = Four-wire receive voltage

Figure 40

at 200 Hz are not uncommon for low-inductance line couplers. The high-end response of these transformers is relatively flat to within 0.5dB at frequencies up to 4kHz (based on a 1kHz reference frequency). Insertion losses may be high, perhaps as much as 3dB, as a result of DCRs over 100 ohms and core losses in the range of a few thousand ohms.

Dry couplers are similar to bridging transformers except they have lower magnetizing inductance and correspondingly less resistance. Typical inductance values may range from 0.5 henries to several henries. Winding resistance values may range from a few ohms to perhaps 100 ohms. Frequency response is typically ±1dB from 200 to 4000 Hz. Insertion losses for these couplers may range from 0.5 to perhaps 2dB.

Hybrid couplers

Hybrid couplers are designed to combine the transmit and receive signals for transmission onto a single wire pair. The combining process of a hybrid is designed to prevent the transmit signal from appearing at the receive port causing oscillations known as “singing”. Figure 40 shows a typical two-transformer hybrid arrangement.

A single transformer may also be configured as a hybrid, although performance will not be as good as a two-transformer hybrid for a given transformer size. A typical single-transformer hybrid is shown in Figure 41.

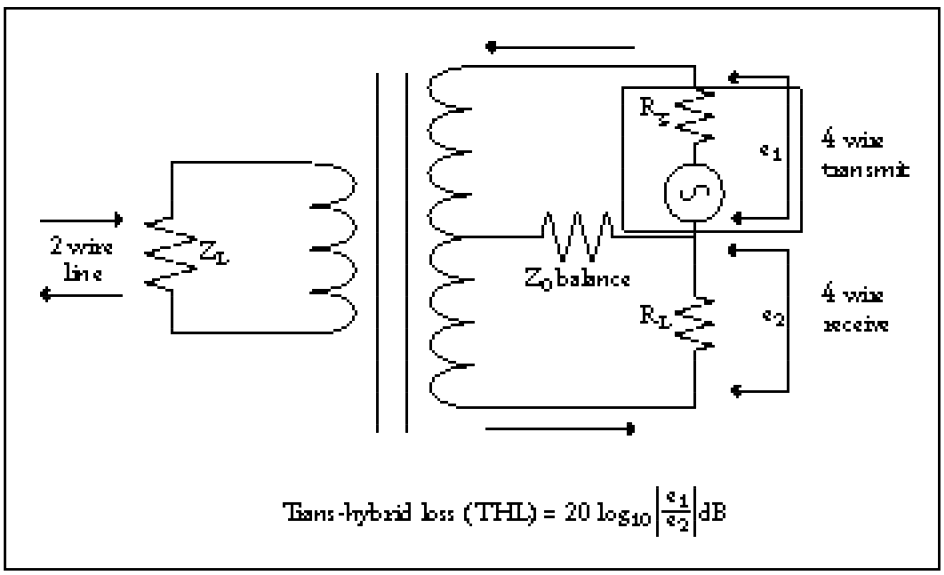


Figure 41

Hybrid transformers were once very popular as a means of providing the two-wire to four-wire conversion required in telephone circuits. Since the late 1970s this function has been fulfilled through the use of operational amplifiers (op-amps) or a number of other specialty circuits.

Hybrid transformers may be wet or dry depending on the application. As with standard couplers, dry hybrid transformers perform better than their wet counterparts since they have more magnetizing inductance. For more information on hybrid transformers, see the Midcom web site,

http://www.midcom-inc.com/technology/terms/thloss.htm

High-fidelity

Microphone

Microphone transformers are used to provide a balanced-to-unbalanced (balun) conversion or to match the impedance of a microphone to that of an amplifier. Such transformers typically have bandwidths from 20 to 20,000 Hz with ±1dB or flatter response. Faraday shields and mu-metal enclosures are used to prevent noise pickup. Transformers with high-turns ratios are subject to mechanically-induced noise pickup called *microphonics* and may be encapsulated with paraffin or similar soft wax to dampen the effects of mechanical shock and vibration. Good balance is essential to remove common-mode interference, a.k.a. "hum" due to the wide dynamic range afforded many audio systems. Typical common-mode rejection requirements may exceed 80 dB and have dynamic ranges in excess of 70dB. Common uses include matching the relatively low 2K ohm output impedance of a microphone to an amplifier's much higher line input impedance of 10K ohms. Studios commonly use the three terminal "XLR" type of connector which is a balanced connection method with a terminal for a center tap. A separate ground terminal, tied to the XLR connector's case is almost always present. The center tap may be used to phantom-feed a small amount of current for powering a pre-amp or active "condenser" microphone. The feeding current is typically supplied via a 48V supply (shades of telecom!). Distortion is also critical for transformers used in recording or broadcasting studios. It is not uncommon that distortion must be controlled to less than 0.05% (-66 dB) at the highest drive level. For more information on High-Fidelity transformers, refer to the excellent Frequently Asked Question (FAQ) list for the Usenet newsgroup rec.audio.pro, which may be found at

http://www.cs.ruu.nl/wais/html/na-dir/AudioFAQ/pr0-audio-faq.html

Impedance matching

Many applications call for impedance matching, although the three main reasons an impedance match is desirable are to minimize reflections, to conserve power and to preserve signal dynamic range. Impedance matching does no good in systems where the benefits do not outweigh the costs. In fact, a lossy, resistor-based network can many times provide a better wide-band match to two different impedances than can a transformer-based solution. (While we'd love to sell everybody a transformer, there are times when they just aren't needed.) For more information on resistive matching networks, refer to Donald Fink's *Electronic Engineer's Handbook*.

Power

One of the most prevalent uses of a transformer is that of providing power to a circuit at a voltage different from that which is readily available. Many power transformers also provide isolation. There is so much information on power transformers that all I can hope to provide here are some general guidelines and perhaps a few caveats that might otherwise not be readily apparent.

Linear

Linear power transformers didn't need the name "linear" until the advent of "switchmode" technology. Until switchmode, the only common power transformer technology was linear. The choice of the term "linear" was made as a direct counter to the method of switching, as we will see later.

- 50/60 Hz

  Transformers for 50 and 60 Hz were almost assuredly the first large-scale commercial transformers manufactured. This makes transformers and electrical motors two of the longest-lived components in electrical history. The applications for 50 and 60 hertz transformers are so large that even a partial list would fill several pages. If a transformer will not be used in a country that has 50 Hz power, save yourself some money by specifying 60 Hz only. Transformers that must operate at 50 Hz will be larger and more expensive than their 60 Hz counterparts (remember the flux density formula?) so if you don't need it, don't ask for it. (Aircraft uses 400 Hz power to make the magnetics even smaller and lighter. If you listen closely when you're on a jet that is still on the ground with its engines off, you can hear the much higher frequency of the power distribution system)

- Autotransformer

  If a transformer is not required to provide isolation, the primary and secondary can share some of their turns (neat trick, huh?). Sharing windings can provide a size and cost savings. Autotransformers are commonly used to provide adjustable voltage when line voltages aren't exactly the desired value. In days of old, televisions weren't built to handle wide ranges of line voltages. In rural areas, the power lines may have provided only 90V when the TV was expecting 117V. In this case, the TV picture didn't fill the screen properly and an autotransformer was used to bring the voltage back into spec. This was done by connecting the 90V line to an autotransformer's common and primary tap terminals with the TV set connected to the common and a secondary terminal with 30% more turns than the primary. Figure 42 shows a typical configuration.

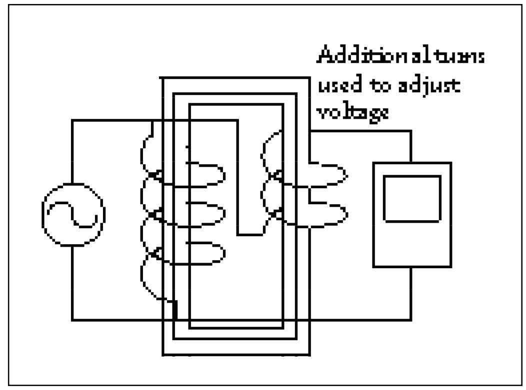


Figure 42

Switchmode

Switchmode power uses an approach different from linear in that the incoming energy is switched on and off at a rate typically much faster than 50 or 60 hertz. Frequencies up to one megahertz are commonly used. As of this writing, 100kHz to 500 kHz is the range where probably 80 percent of all switchmode supplies operate. Switchmode supplies are more complex than their linear counterparts, and only a few years ago were cost-prohibitive for applications less than 30 watts. Recent developments by switchmode IC makers, in conjunction with suppliers of magnetic components, have resulted in the moving of the dividing line to a power level around 10 watts.

A typical switchmode power supply is shown in Figure 43. This particular arrangement is called an off-line switchmode supply because it is connected directly to the power line (which really doesn't make much sense because it should be called an on-line supply!). The circuit consists of a rectifier/filter block that converts the 117VAC into about 165VDC. This DC voltage is 'chopped' by the switch S at a rate

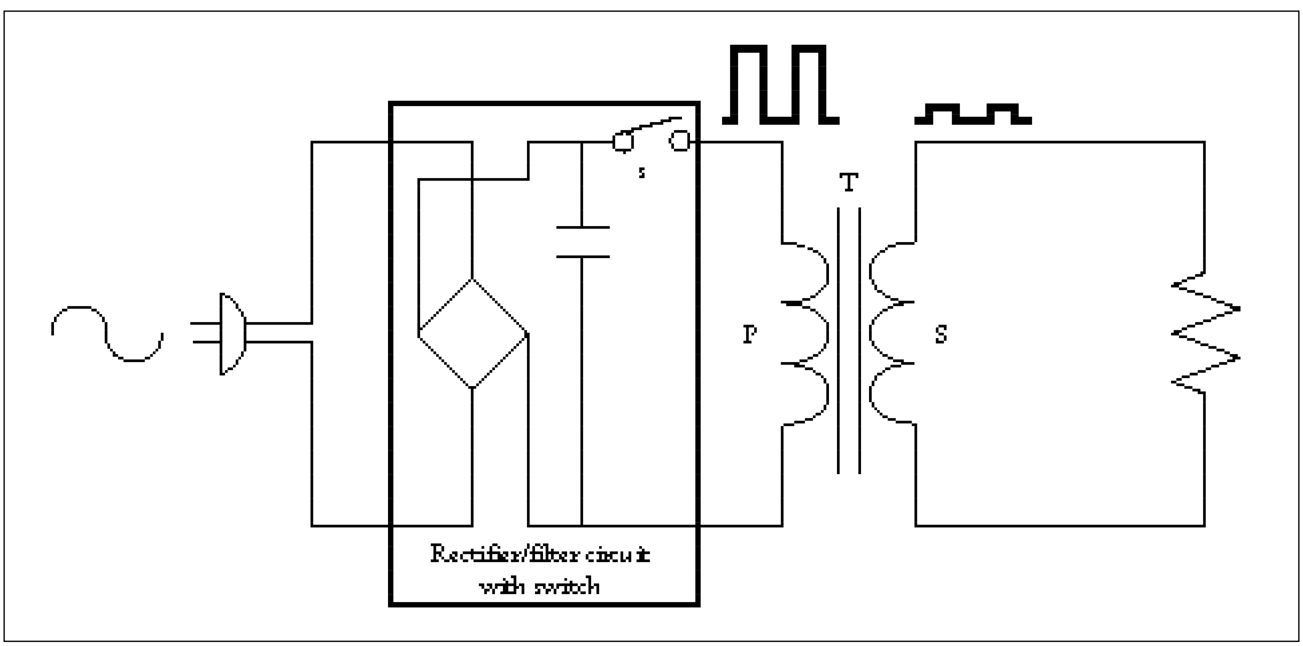


Figure 43

appropriate for use with transformer, T. Losses are reduced by sending the power across the transformer in a pulsed mode. Since a switch is either an open circuit or a short, it can't (in theory) dissipate any energy. The switching frequency is chosen to be high enough that only a small number of turns are required to support the flux density, thus the resistance of the windings can be quite low. It is for these reasons that switchmode supplies are much more efficient than their linear counterparts, given an equivalent size constraint.

Topologies

The flyback topology is by far the most popular and cost-effective topology for switchmode power applications up to 100 watts. Flyback supplies are simple to design and minimize the demands on the magnetic components. They also require a minimum of support circuit components to make them run.

Other topologies include forward converter, buck, boost, push-pull, half-bridge and full-bridge. Contact Midcom Engineering for more information on power supply topologies.

Proximity effect

A phenomenon known as proximity effect can cause a transformer's winding to lose some of its current-carrying capacity. If the winding is close to another winding that is carrying a high current, that current can cause a redistribution of current in the adjacent winding. Magnetic flux linkages from the high-current winding will effectively push the current toward one side of the other winding's conductor, rendering all but the portion of the side closest to the other conductor's cross-section useless. The proximity effect may require a transformer designer to choose a wire size larger than would normally be required for a given current since the flux linkages effectively narrow the conductor's cross-sectional area.

Power factor correction

For best efficiency, it is desired that all loads appear to be purely resistive. If a load is instead reactive, either capacitively or inductively, it will cause a phase shift and cause a mismatch between source and load which leads to a lack of perfect power transfer. Power Factor is the term given to the amount shift in the phase angle between the current and the voltage entering the load. Motors tend to introduce an inductively-reactive phase shift while most switchmode supplies introduce a capacitive phase shift. (Remember ELI the ICE man – voltage, E leads current, I in an inductive "L", circuit; and current, I leads voltage, E, in a "C" or capacitive circuit). The Power Factor, or PF, is the cosine of the angle between the voltage and current. Power factor correction circuits work by forcing a current to compensate for phase lag.

Inductors

You've heard of inductors, and you've heard of chokes. Do you know what the difference is between them? The price! (An 'inductor' just sounds more expensive than a plain, old 'choke'.) Inductors for switchmode applications must have wire diameter suitable for carrying the peak current, not just the average or RMS current (same is true of switchmode transformers). The inductors may also require shielding if their core structure is "open" such as those built on drum or rod cores. A better approach is to keep them away from things that are affected by magnetic fields. CRTs and other open inductors are things to keep away from high-current inductors. The distributed capacitance of an inductor can cause it to have a low self-resonant frequency (SRF). This can render it useless for high-frequency applications. Most inductor manufacturers will provide you with a minimum SRF for their products. You can measure it very easily with an LCR bridge by sweeping the frequency and noting where the impedance of the inductor turns the corner from inductive to capacitive. You can also measure the magnitude of an inductor's impedance and note the frequency at which its impedance is lowest.

Tunable

Most tunable inductors have ferrite cores. Tunable inductors can be tricky to design. There are at least two ways of making an inductor tunable. One is to use a threaded plastic insert which contains a small piece of ferrite which slides into a gap in the core halves as the insert is rotated. Another method involves a special shape, usually a slant cut, to the core halves center posts which causes them to reduce the air gap between the posts as one core half is rotated with respect to the other. This second method is less popular than the first for reasons of convenience.

The tricky part about designing a tunable inductor is tolerance buildup. The cores have an inductance tolerance, usually ±3% and the threaded insert also has a minimum and maximum shift that influences the permeability of the air gap. Inserts are commonly specified by the amount of percent shift they will cause a core of a given Al value. A third variable to consider is that turns are wound in integer amounts, so if the required inductance value is far from what an integer number of turns can provide, the result is a lessening of the amount of overall shift available. When the design is complete, and adjuster rated for a 15% shift might only provide a guaranteed tunable inductance range of 8%.

Test-fixture transformers

Return loss

A transformer can be used to construct a return loss test fixture. Midcom's standard test method, 999-2318, describes the use of our 671-0023 in a dandy return loss test fixture. Rather than reproduce it here, I will give you the web address so you can view it yourself:

http://www.midcom-inc.com/technology/terms/lossmezr.html

The transformer is used in a hybrid configuration to develop to equal voltages across the halves of its secondary winding. The center tap of the transformer provides a half-way point for the voltage. The entire secondary voltage is applied to a series connection of the device under test and a reference impedance network. The voltage from the center tap to the connection point between the DUT and the reference network is compared to the voltage across the reference network only. The log of the ratio of these two voltages yields the return loss in dB (with a multiplier of 20 to make the math work out properly).

Longitudinal balance

Longitudinal Balance transformers must be constructed to provide at least 20 dB better balance than the devices they are used to test. The 20 dB factor is another way of saying that the measuring equipment must be 100 times better than the device being measured. Faraday shields and careful winding schemes are employed to assure balance. Even so, a longitudinal balance test transformer is typically valid for only three to perhaps four decades of frequency. Midcom's 671-0323 and 671-5767 are used for voiceband and subrate applications respectively. The 671-0323 is good for 15 Hz to 15 kHz while the 671-5767 spans a range from 10kHz to about 2MHz. Midcom test specification 999-2286 provides a circuit and test fixture calibration method for longitudinal balance test fixtures using those transformers. The Midcom web site also contains information on these fixtures at

http://www.midcom-inc.com/technology/terms/balance.html

Balanced-feed inductors

Sometimes it is necessary to feed DC to a device under test. In the case of longitudinal balance, care must be taken that the DC feed supply does not unbalance the test in progress. In this situation a balanced feed inductor is almost mandatory (although a chemical battery with appropriate resistors can be used if you're willing to discard them when they run out). The balanced feed inductor must be carefully balanced, again to a level 20 dB better than the measurement to be taken. Since the inductor must be gapped to prevent it from saturating from the DC current, the position of the gap can be crucial with respect to balance. In this case it is best that each inductor be 'tuned' for best balance by adjusting the gap slightly within the coil. This is particularly true if the balanced feed inductor incorporates a two-section coil such as Midcom's 671-4130, which may be viewed at:

http://www.midcom-inc.com/products/other/671-4130.html

Afterword

This Technical Note has been a long, long time in the making. Even so, I plan to revise it as needed to keep it current. If you have comments, suggestions or questions regarding its content, please send them to me at the address below.

Thanks for your attention!

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Glossary

bobbin

A device used to facilitate the winding of a coil by containing the turns of a conductor during the winding process. Many bobbins have terminals onto which the conductor ends are connected.

clearance

The line-of-sight distance between two conductors along an insulating medium.

closed path

A homogeneous magnetic core structure is said to provide a closed path to a magnetic field. We stretch this just a bit to say that even core structures which contain a small air gap are also closed path structures. Once the air gap becomes large with respect to the core's magnetic path length, the structure is said to have an open path. A common example of a closed path structures is the uncut toroidal core. An example of an open path is the common bar magnet. The drum core (or bobbin core) is another example of an open path device.

comparative tracking index (CTI)

Insulating materials are graded by their ability to resist the effects of breakdown to continuously-applied voltage. The name for the rating is known as the Comparative Tracking Index, or CTI. CTI is tested by applying a voltage to two conductors attached to a test specimen of the insulating material, to which a contaminant such as water or an ammonia mixture has been applied. The CTI rating is the voltage at which carbon traces or "tracks" begin to appear. A typical plastic may have a CTI rating of 600V.

coil

A conducting filament arranged in a helix or spiral around a magnetic core. Most coils are wound with magnetwire, but other conductor materials may also be used. Coils may be wound on coil formers, or may be free-standing, although this last arrangement is only used when the number of coil turns is small and the conductor is stiff enought to allow the coil to hold itself together.

coil former

See *bobbin.*

core

The magnetic structure around which the coil is located. The core need is usually consists of a magnetic material, but "air-core" coils are also common which contain no material at all.

creepage

The shortest distance between two conductors separated by an insulating medium defines the minimum creepage distance between those two conductors. Safety agencies are only concerned with the minimum creepage distance between two conductors as it is over that distance that high voltage breakdown is most likely to occur.

DC resistance (DCR)

The resistance of a coil winding when a steady-state (zero hertz) or DC voltage is applied. See also: *AC resistance*.

dot convention

Where dots meet to hobnob with other dots. Also the notation system used to describe the polarity of a transformer's windings. A current flowing into the "dotted" terminal of a transformer winding will cause a current to flow out of the "dotted" terminal of another winding. The dots are usually shown on the schematic drawing of the transformer, but are sometimes applied to the transformers themselves.

equal area theorem

The amount of droop of a pulse waveform versus the ideal waveform desribes energy that magnetizes the transformer core. That energy must be conserved and presents itself at the transformer's output winding in the form of backswing. The equal-area theorem is based on the concept that a transformer cannot carry DC. As a result, the amount of pulse area above the zero axis must equal the amount below the axis. Sort of a magnetic equivalent to "water seeks its own level."

finish terminal

The wire or terminal to which the end of a winding is connected. See also: *start terminal*.

fringing

Fringing describes the effect that causes a magnetic field to bow outward from a gap in the core structure. Fringing may cause a conductor turn to be magnetically absent from a coil if that turn is near enough to the gap.

Galvanic isolation

When two conductors are insulated from one another they are said to be galvanically isolated. Use of the term galvanic isolation is a way of expressing the extent to which we can count on this form of isolation. It is possible to isolate two conductors with high-value resistors, but this form of isolation does not meet the definition of galvanic isolation. Galvanic isolation should be reserved to define situations where electron flow is essentially zero between two conductors of differing potential.

induction coil

Older term for transformer, in fact the name by which transformers were known for the first hundred or so years of their existence. The term is still used in some texts on telephony.

magnetwire

A special class of wire characterized by a very thin insulation coating which allows very tight packing of wire into a small volume. Many magnetwires are solderable, meaning their insulation coating may be removed by a soldering process, thus revealing the bare conductor underneath. Magnetwire is available in various cross-sectional shapes, such as square and rectangular, but by far the most common shape is circular.

permeability

The characteristic of a core material that describes its ability to store the energy of a magnetic field. The permeability of free space (a volume containing no magnetic material) in MKS units is 4(X10-7 henries/meter. The permeability of free space in CGS is unity. See also: *relative permeability, effective permeability.* Another way to define permeability is the ratio of magnetic flux density, B, in the volume of a given material to the flux density, H, that would appear if that same volume consisted solely of free space.

permeability, effective

The base permeability of a core material may be changed by introducing a gap into the core's magnetic path. If this is done, the overall permeability of the core is reduced and we call this new permeability the effective permeability.

permeability, relative

It is easier to compare various core materials by expressing their permeability as a ratio over that of free space. Given that free space has a permeability of 4(X10-7 henries/meter, a core material 10,000 times more permeable would thus have a permeability of 4(X10-3 henries/meter (MKS units). Rather than rate it in that way, we refer to the material's relative permeability, which in this case is 10,000. Since the permeability of free space in CGS is unity, relative permeability is identical to the material's base permeability when dealing with the CGS system.

polarity

Since a transformer's windings can be completely isolated form one another, the transformer may be used to reverse the polarity of a given signal. See dot convention for the method used to denote this ability for a given transformer.

pollution degree

Electrical equipment may be used in various environments. Safety agencies have classified these environments according to their likelyhood of having internal contaminants which could impair the function of the insulators used in the equipment. The classifications are known as pollution degrees, and in EN60950 there are three. Pollution degree I applies to non-vented, hermetically sealed devices. Pollution degree II refers to electrical equipment with vents or other means of allowing clean air to enter and exit. Equipment in pollution degree III have internal conducting contaminants or are intended for use in environments which include such contaminants.

primary

The winding of a transformer usually connected to the power mains or other transmission line. The term primary is really just a way to name a winding and doesn't confer any particular meaning with respect to the order of the windings as they were created, nor how the winding is used in the circuit.

secondary

The winding of a transformer which is most likely to be connected to the "circuitry" (non-transmission line) side of the transformer.

skin depth

The skin depth of a conductor describes the effective cross-sectional area available to an alternating current as a result of the skin effect. The skin depth is the point at which the ac field amplitude of a conductor has dropped to 1/e or 36.8% of its surface volume.

skin effect

When an alternating current passes through a conductor, it causes eddy currents magnetic fields that redistribute the current flow in the conductor. The redistribution of current increases the resistance of the center of the conductor to a point where the effective cross-sectional area of the conductor is reduced to a fraction of its DC value. At very high frequencies almost no current flows at the center of the conductor such that a tube will have almost the same current-carrying capacity as a solid conductor. See skin depth for a mathematical description of this effect.

ìstartî terminal

When a transformer winding is first created, the wire corresponding to the beginning of that winding is called the start. This name also applies to the terminal where the start winding is connected. If the rotation direction of all windings on a given core structure are the same, all ëstart' terminals will have the same electrical polarity. Similarly, all ëfinish' terminals will have the same polarity. Many manufacturers apply the polarity dots to the start terminals, just for the sake of consistency and ease in proofreading winding prints. This rule only applies if the winding direction about the core is identical for all windings. The polarity dot is the true indicator of winding polarity. The physical starts and finishes may not correlate to winding polarity if winding direction was changed during the course of transformer construction.

switchmode power conversion

The process of converting electrical power from one voltage to another by first converting the power source to DC, then rapidly turning the DC on an off, thus creating a high-frequency AC. The high-frequency AC is put through a transformer to change its voltage to the desired amount, then usually rectified and filtered to provide the desired output level.

References

William Dull, Dr. A. Kusko and Thorleif Knutrud, “Pulse and Trigger Transformers - Performance Dictates their Specs”, EDN, August 20, 1976

Nathan Grossner, “Transformers for Electronic Circuits” McGraw-Hill Book Company, ISBN 0-07-024979-2