Magnetics Design for Switching Power Supplies

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Section 1 Introduction and Basic Magnetics

Introduction

Experienced SwitchMode Power Supply designers know that SMPS success or failure depends heavily on the proper design and implementation of the magnetic components. Parasitic elements inherent in high frequency transformers or inductors cause a variety of circuit problems including: high losses, high voltage spikes necessitating snubbers or clamps, poor cross regulation between multiple outputs, noise coupling to input or output, restricted duty cycle range, etc. Figure 1 represents a simplified equivalent circuit of a two-output forward converter power transformer, showing leakage inductances, core characteristics including mutual inductance, dc hysteresis and saturation, core eddy current loss resistance, and winding distributed capacitance, all of which affect SMPS performance.

With rare exception, schools of engineering provide very little instruction in practical magnetics relevant to switching power supply applications. As a result, magnetic component design is usually delegated to a self-taught expert in this "black art". There are many aspects in the design of practical, manufacturable, low cost magnetic devices that unquestionably benefit from years of experience in this field. However, the magnetics expert is unlikely to be sufficiently aware of the SMPS circuit problems caused by the various parasitic elements and the impact of the specific circuit locations of these elements. This often results in poor decisions in the magnetic component design.

This collection of topics on magnetics is intended to give the SMPS designer the confidence and the ability to: (1) Develop a reasonably accurate electrical circuit model of any magnetic device, to enable prediction of circuit performance, (2) Relate the electrical circuit model to the magnetic device structure, thus providing the insight needed to achieve an optimized design, (3) Collaborate effectively with experts in magnetics design, and possibly (4) Become a "magnetics expert" in his own right.

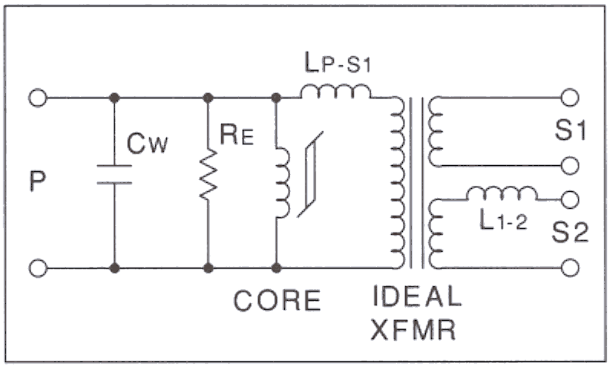


Figure 1-1 Transformer Equivalent Circuit

Obstacles to learning magnetics design

In addition to the lack of instruction in practical magnetics mentioned above, there are several other problems that make it difficult for the SMPS designer to feel "at home" in the magnetics realm:

- *Archaic concepts and practices.* Our great-grandparents probably had a better understanding of practical magnetics than we do today. Unfortunately, in an era when computation was difficult, our ancestors developed concepts intended to minimize computation (such as "circular mils") and other shortcuts, still in use, which make the subject of magnetics more complex and confusing to us today. Ancient design equations intended for sinusoidal waveforms (and not clearly defined as such) are often incorrectly applied to rectangular SMPS waveforms.
- *The CGS system of units.* Magnetics design relationships expressed in the modern SI system of units (rationalized MKS) are much simpler than in the archaic CGS system. SI equations are much easier to understand and remember. Unfortunately, in the U.S., core and magnet wire data is

usually published in the old-fashioned CGS system, with dimensions often in inches, requiring data conversion to apply to the SI system.

- *Energy/time vs. power.* Circuit designers are comfortable in the electrical realm of Volts and Amperes. On a Volt/Ampere plot, area represents power. Time is not directly involved.

  But on a plot of magnetic flux density, B, vs. field intensity, H, area represents *energy.* Time is always required to change flux density, because an energy change must take place. A change in flux density, ΔB, within a winding equates to the *integral* volt-seconds per turn across the winding (Faraday's Law). Time is definitely involved. This concept takes some getting used to.

An appeal to suppliers of core materials and wire: Old-time magnetics designers in the U.S. are acclimated to the CGS system, and may prefer magnetics data expressed in Gauss and Oersteds. But newcomers to magnetics design, as well as experienced designers outside the U.S. prefer the internationally accepted SI system – Tesla and Ampere-Turns. Also, cores and wire dimensional data should *definitely* be provided in metric units.

Core losses are usually characterized as a function of frequency using sinusoidal current-driven waveforms. This is erroneous and misleading for high frequency SMPS applications. High frequency core losses are primarily caused by eddy currents, which depend on rate of flux change, *not* on frequency per se. (At a fixed switching frequency, higher V_{IN} with short duty cycle results in higher loss.) It would be most helpful if materials intended primarily for SMPS applications were characterized using rectangular voltage-driven waveforms, with examples shown of core loss and minor hysteresis loops under these conditions.

Magnetic Field Relationships

Understanding the rules that govern the magnetic field is extremely valuable in many aspects of switching power supply design, especially in minimizing parasitics in circuit wiring, as well as in magnetic device design.

Figure 1-2 shows the field surrounding two parallel conductors, each carrying the same current but in opposite directions, i.e. a pair of wires connecting an electrical source to a load.

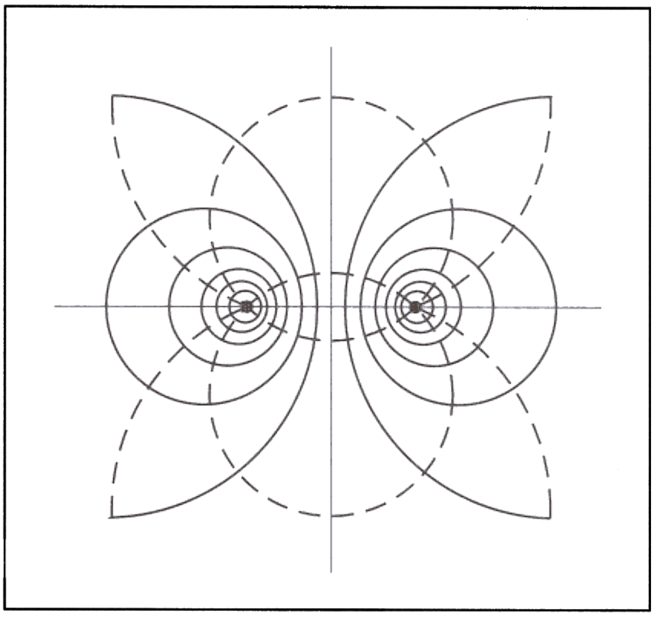


Figure. 1-2 Field Around Conductor Pair

The solid lines represent magnetic *flux*, while the dash lines represent an edge view of the *magnetic field equipotential surfaces.* Each wire has an individual field, symmetrical and radial, with field intensity diminishing in inverse proportion to the distance from the conductor. These two fields are of equal magnitude but opposite in polarity because the currents that generate the fields are in opposite directions. As shown in Fig. 1-2, the two fields, by superposition, reinforce each other in the region between the two wires, but elsewhere they tend to cancel, especially at a distance from the wires where the opposing field intensities become nearly equal.

Figure 1-3 shows the field associated with a simple air cored winding. The individual fields from each wire combine to produce a highly concentrated and fairly linear field within the winding. Outside the winding, the field diverges and weakens. Although the stored energy density is high within the winding, considerable energy is stored in the weaker field outside the winding because the volume extends to infinity.

A magnetic field cannot be blocked by "insulating" it from its surroundings – magnetic "insulation" does not exist. However, the magnetic field *can* be short-circuited – by placing the coil of Fig 1-3 inside

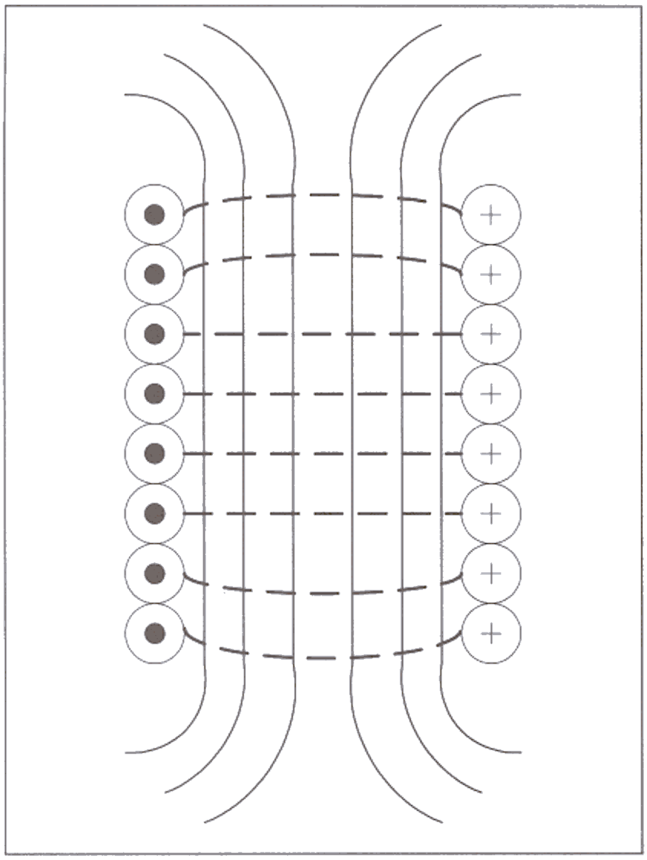


Figure. 1-3 Air-Core Solenoid

a box of high permeability magnetic material, which provides an easy path for the return flux, shielding the coil from the environment external to the box

Important Magnetic Field Principles

- The total magnetic field integrated around any closed path equals the total ampere-turns enclosed by that path (Ampere's Law)
- Magnetic field equipotentials are *surfaces*, not lines. (Alternatively, field intensity can be represented as a vector normal to the surface.)
- Magnetic field equipotential surfaces are *bounded* and *terminated* by the current which generates the field. They are not closed surfaces, as with electric field equipotentials.
- Flux is in the form of *lines*, not surfaces. (Flux can also be represented as a vector.)
- Flux lines are *always closed loops* – they never begin or end. In any arbitrary volume, the number of flux lines entering must equal the number leaving, regardless of the contents of that volume.
- Flux lines are always normal to the magnetic field equipotential surfaces.
- At any location, flux density is always proportional to field intensity: $B = \mu H$

Conservation of energy: At any moment of time, the magnetic field and the current flow distribute themselves so as to minimize the energy taken from the source. If alternative current paths exist, current will initially flow in the path(s) resulting in minimum stored energy, but in the long term, the current flow redistributes so as to minimize I^2R loss. Reference (R2) has more on this subject.

Transformation of Axes

SMPS circuit designers are obviously interested in the electrical characteristics of the magnetic device as seen at the device terminals.

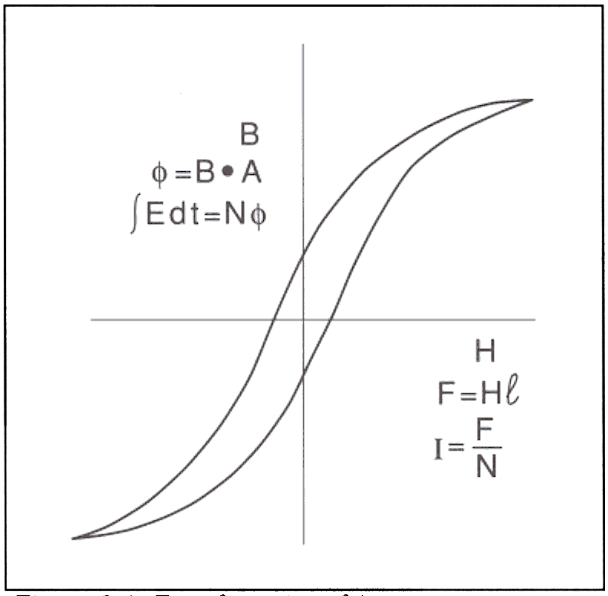


Figure. 1-4 Transformation of Axes

Figure 1-4 shows how the horizontal and vertical axes scale factors can be altered so that the *B-H* characteristic (defining a core *material*) is translated into a ϕ vs. $\mathcal{F}$ (mmf) characteristic (defining a *specific core* with magnetic area A_e and path length ℓ_e). Transforming the axes once again using Faraday's Law and Ampere's Law, the same curve now represents the equivalent electrical characteristics of that core when wound with N turns, $\int Edt$ vs. I.

Note that the slope of the *B-H* characteristic is *permeability*, the slope of the ϕ vs. $\mathcal{F}$ characteristic is *permeance*, while the slope of the $\int Edt$ vs. I characteristic is *inductance*.

Systems of Units

The internationally accepted SI system of units (Système International d'Unités) is a rationalized system, in which permeability, $\mu = \mu_0 \cdot \mu_r$ (μ_0 is the absolute permeability of free space or nonmagnetic material = $4\pi \cdot 10^{-7}$; μ_r is the relative permeability of a magnetic material). In the unrationalized CGS system, $\mu_0 = 1$, therefore μ_0 is omitted from CGS equations so that $\mu = \mu_r$. But the rationalization constant μ_0 doesn't just disappear in the CGS system – instead, portions of this constant show up in all the CGS equations, complicating them and making them more difficult to intuitively grasp. In the SI system, all of the "garbage" is gathered into μ_0, thereby simplifying the SI equations.

The equations below are given in both systems – SI and CGS. It is suggested that beginners in magnetics design stick to the SI equations and ignore the CGS system until completely comfortable with the principles involved. Then, it may be helpful to use the CGS system when working with magnetics data expressed in CGS units, rather than convert the units.

Table I
Magnetic Parameters and Conversion Factors

		SI	CGS	CGS to SI
FLUX DENSITY	B	Tesla	Gauss	10^{-4}
FIELD INTENSITY	H	A-T/m	Oersted	$1000/4\pi$
PERMEABILITY (space)	μ_0	$4\pi \cdot 10^{-7}$	1	$4\pi \cdot 10^{-7}$
PERMEABILITY (relative)	μ_r			
AREA (Core Window)	A_c, A_w	m^2	cm^2	10^{-4}
LENGTH (Core, Gap)	ℓ_e, ℓ_g	m	cm	10^{-2}
TOTAL FLUX = $\int BdA$	ϕ	Weber	Maxwell	10^{-8}
TOTAL FIELD = $\oint Hd\ell$	$\mathcal{F}$,mmf	A-T	Gilbert	$10/4\pi$
RELUCTANCE = $\mathcal{F}/\phi$	$\mathcal{R}$	A-T/Wb	Gb/Mx	$10^9/4\pi$
PERMEANCE = $1/\mathcal{R}$	$\mathcal{P}$			$4\pi \cdot 10^{-9}$
INDUCTANCE = $\mathcal{P} \cdot N^2$ (SI)	L	Henry	(Henry)	
ENERGY	W	Joule	Erg	10^{-7}

Ampere's Law and Faraday's Law jointly govern the important relationship between the magnetic elements and the equivalent electrical circuit as seen across the windings.

Ampere's Law

SI:

$$F = \oint Hd\ell = NI \approx H\ell \quad \text{A-T}$$

$$H \approx NI/\ell \qquad \text{A-T/m (1)}$$

$$F = \oint Hd\ell = .4\pi NI \approx H\ell \qquad \text{Gilberts}$$

$$H \approx .4\pi NI/\ell \qquad \text{Oersteds (1a)}$$

Amperes Law states that the total magnetic force, $\mathcal{F}$, along a closed path is proportional to the ampere-turns in a winding or windings linked to that path, i.e., that the path passes through. In the SI system, the units of magnetic force are expressed in ampere-turns. When the field *intensity* H varies along the path, H must be integrated along the path length. Fortunately, the simplified form shown in Eq. 1 and 1a can be used in most situations.

Faraday's Law

SI:

$$\frac{d\phi}{dt} = -\frac{E}{N}; \quad \Delta\phi = \frac{1}{N}\int Edt \qquad \text{Weber (2)}$$

$$\frac{d\phi}{dt} = -\frac{E}{N} \times 10^8 \quad \Delta\phi = \frac{10^8}{N}\int Edt \quad \text{Maxwell (2a)}$$

Faraday's Law equates the flux *rate of change* through a winding to the volts/turn applied to the winding. Thus, the flux change is proportional to the integral volt-seconds per turn (directly equal in the SI system). Faraday's Law operates bilaterally – that is, if 2.5 volts/turn is applied to winding A, the flux through A will change by 2.5 Webers/second. If a second winding, B, is linked to all of the flux produced by winding A, then 2.5 Volts/turn will be *induced* in B.

Faraday's Law makes it clear that flux cannot change instantaneously. Any flux change requires time and usually a change in energy. Time is required to move along the ϕ or B axis, which is more obvious considering the electrical equivalent scale dimensions are Volt-seconds.

Note that all flux lines follow a closed loop path. Flux lines have no beginning or end.

Energy

SI:

$$W/m^3 = \int HdB \approx \frac{1}{2}BH$$

$$W = \int Vol \cdot HdB = \int I \cdot Edt \qquad \text{Joules} \quad (3)$$

CGS:

$$W = \int Vol \cdot HdB = \int I \cdot Edt \cdot 10^{-7} \qquad \text{Ergs} \quad (3a)$$

Energy put into and removed from the magnetic system can be determined by integrating the area between the characteristic and the *vertical* axis (B, ϕ, $\int Edt$) on the energy plane. Energy must be integrated over time, which is a factor on the vertical axis, not the horizontal.

It is much easier to understand this process by using the electrical equivalent axes, Volt-seconds and Amperes. Referring to Fig. 1-5, from point A to B, energy from the external circuit is put into the magnetic system, as shown by the shaded area between A-B and the vertical axis. From B to C, magnetically stored energy is returned to the electrical circuit. The difference between the energy put in and taken out is hysteresis loss, the area between the two curves. At C, the magnetically stored energy is zero.

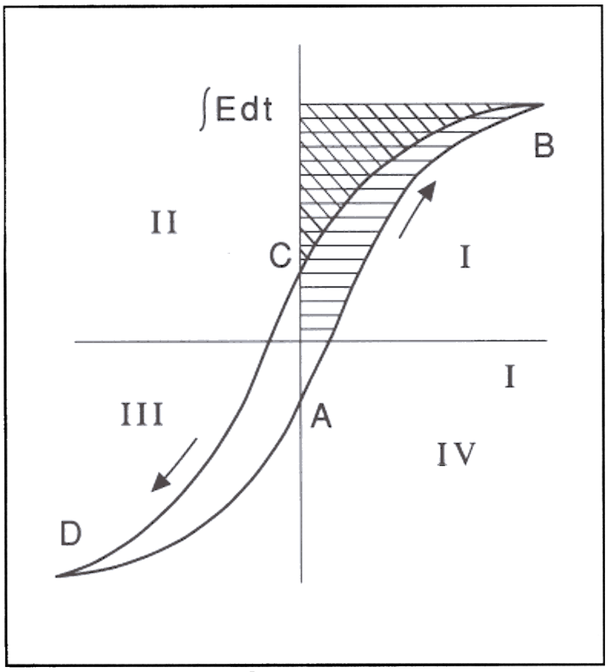


Figure 1-5 Energy Plane

From C to D, energy is put into the system. From D back to A energy is returned to the electrical circuit. The area between the curves is loss. At A, the remaining stored energy is zero.

A positive energy sign indicates energy put in; a negative sign indicates energy returning to the external circuit. From A to B, voltage and current are both positive, so the energy sign is positive. Although the integrated Volt-seconds are negative at A, *upward movement indicates positive voltage.* From B to C, current is positive but voltage is negative (downward movement). Therefore the energy sign is negative. From C to D, current and voltage are both negative, hence positive energy. From D to A, negative current with positive voltage indicates returning energy.

Permeability

SI:

$$\mu = \mu_0 \mu_r = B/H \qquad \text{Tesla/A-T/m}$$

CGS:

$$\mu = \mu_r = B/H \qquad \text{Gauss/Oersted}$$

Permeability is a measure of a magnetic *material* – the amount of flux which a magnetic field can push through a unit volume of the material. Permeability is roughly analogous to conductivity in the electrical realm.

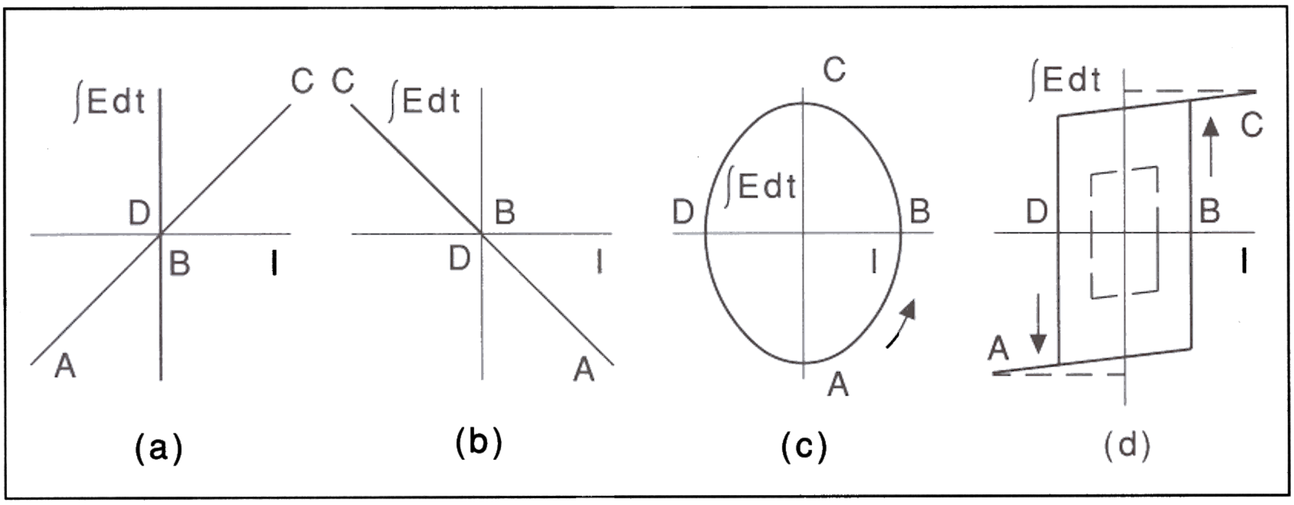


Figure 1-6 – Energy Plane, Sinusoidal Voltage Drive Examples

Permeance, Reluctance

SI:

$$\mathcal{P} = 1/\mathcal{R} = \phi/\mathcal{F} = BA/H\ell$$

Webers/A-T

$$\mathcal{P} = 1/\mathcal{R} = \mu_0\mu_r \bullet A/\ell$$

CGS:

$$\mathcal{P} = 1/\mathcal{R} = \phi/\mathcal{F} = BA/H\ell$$

Maxwells/Gilbert

$$\mathcal{P} = 1/\mathcal{R} = \mu_r \bullet A/\ell$$

When the material characteristic, permeability, is applied to a magnetic element of specific area and length, the result is *permeance*. In the SI system, permeance is equal to the inductance of a single turn.

Reluctance, the reciprocal of permeance, is analogous to resistance in an electrical circuit. (Don't push this analogy too far – reluctance is an energy storage element, whereas resistance is a dissipative element.) Reluctance and permeance can be defined for the entire magnetic device as seen from the electrical terminals, but it is most useful to define the reluctance of specific elements/regions within the device. This enables the construction of a reluctance model – a *magnetic* circuit diagram – which sheds considerable light on the performance of the device and how to improve it. From the reluctance model, using a duality process, a magnetic device can be translated into its equivalent electrical circuit, including parasitic inductances, such as shown in Fig. 1-1.

This will be discussed in a later section.

Inductance

SI:

Henrys (4)

$$L = N^2\mathcal{P} = \mu_0\mu_r N^2 \bullet A/\ell$$

CGS:

Henrys (4a)

$$L = 4\pi\mu_r N^2 \bullet A/\ell \bullet 10^{-9}$$

Inductance has the same value in the SI and CGS systems. In the SI system, inductance is simply the permeance times the number of turns squared.

At Home on the Energy Plane

It is, of course, possible to plot any type of electrical device on the energy plane ∫Edt vs. I. Figure 1-6 shows several different devices – inductors, capacitors, resistors – driven by a sinusoidal voltage waveform. Before looking at Figure 1-7 which shows the waveforms involved, try to identify the devices represented in Fig. 1-6.

The answers:

Fig.1-6 (a) is an air core inductor, ideal and lossless. Current lags the applied voltage, but as shown in Fig. 1-7, the inductor current is in phase with ∫Edt plotted on the vertical axis of the energy plane. As the waveform traverses from B to C, voltage and current are both positive – energy is put into the inductor. from C to D, current is positive but voltage is negative – the same energy previously stored is given back to the circuit.

Fig.1-4 (b) is an ideal, lossless capacitor. Current leads the applied voltage and is therefore 180° out of phase with ∫Edt. From A to B, voltage and current are both positive – energy is put into the capacitor. From B to C, the same energy previously stored is returned to the circuit.

Fig.1-6 (c) is a resistor. Current is in phase with applied voltage, therefore current leads ∫Edt by 90°. Since voltage and current are in phase, their signs are always the same. The energy sign is always positive – energy is always put into the resistor, never returned to the circuit. The entire area within the ellipse represents loss.

Of course, Faraday's Law does not apply to a resistor or capacitor. Therefore the vertical ∫Edt scale for these devices cannot be translated into flux.

Fig.1-6 (d) is an inductor with idealized metal alloy core with low frequency hysteresis, driven into saturation. A tape-wound Permalloy core driven at low frequency (no eddy currents) approaches this characteristic. The area within the characteristic is hysteresis loss. The only energy returned to the circuit is the area of the thin wedges above and below the characteristic. Only when the core saturates, taking on the characteristic of an air core, is any energy stored and returned.

The dashed characteristic shows a minor hysteresis loop occurring with reduced drive, which does not take the core into saturation.

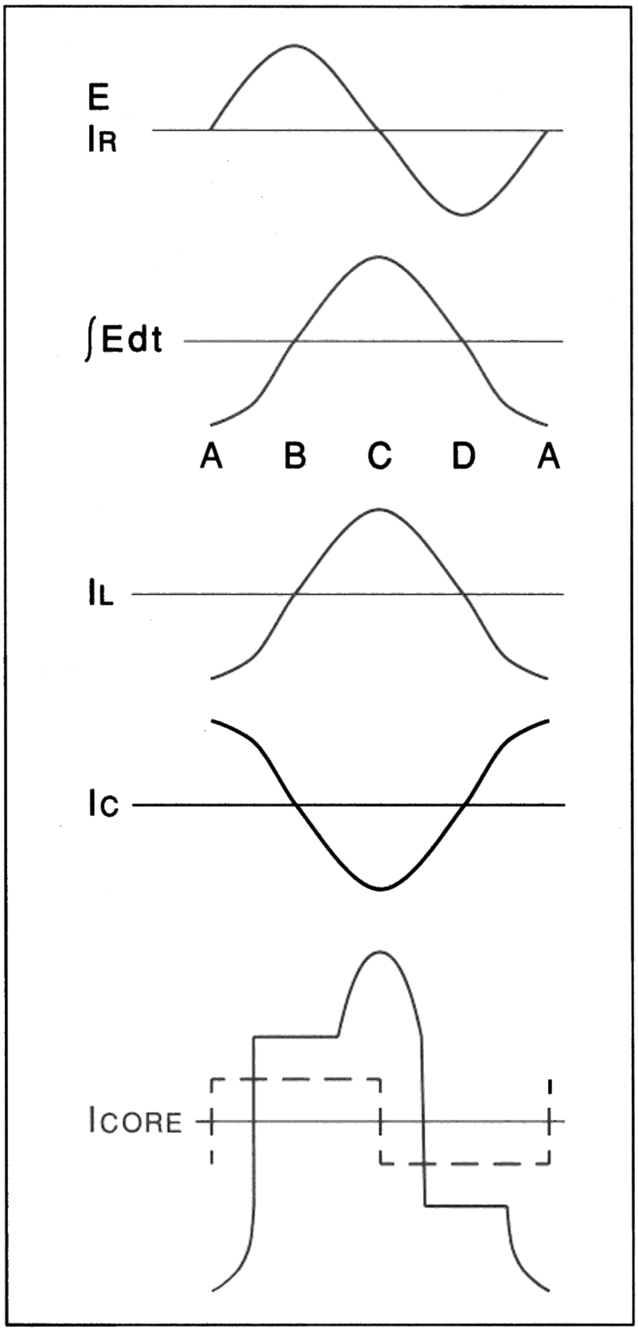


Figure 1-7 Sinusoidal Voltage Drive

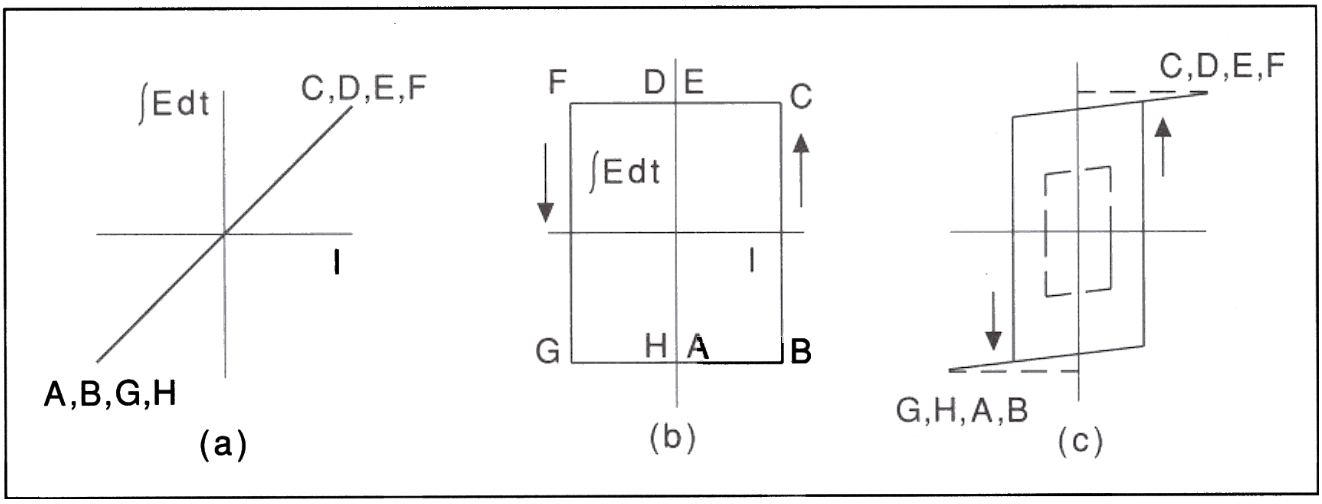


Figure 1-8 – Rectangular Voltage Drive Examples

Rectangular Voltage Drive Waveform

Sinusoidal waveshapes are not relevant in most SMPS applications. Figure 1-8 shows the same devices driven by a symmetrical rectangular voltage waveform (not including the capacitor, which would require an infinite current at the instantaneous voltage transitions). Figure 1-9 shows the corresponding waveforms.

Fig.1-8 (a) is the same lossless air-core inductor as in 1-6 (a). Although the characteristic looks the same as with sinusoidal drive, Fig. 1-9 waveforms reveal that the characteristic dwells at its extremes during times of zero applied voltage.

Fig.1-8 (b) The same resistor as in 1-6 (c) plots as a rectangle rather than an ellipse. The rectangular voltage and current waveforms exist at three distinct levels. From B to C, the voltage and current are both at a constant positive level, while ∫Edt slowly rises. Thus, the current is constant while ∫Edt changes. From C to D, current suddenly collapses to zero, where it dwells until time E, because ∫Edt does not change while the voltage is zero. At time F, the current suddenly changes to its constant negative level, where it remains while ∫Edt slowly drops toward G.

Fig.1-8 (c) The same idealized metal-cored inductor as Fig.1-6 (d) exhibits the same shape on the energy plane although the driving waveshape is quite different.

In fact, with any practical inductor with magnetic core, the low-frequency hysteresis loop (excluding eddy currents) does not change shape radically when frequency *or* voltage *or* the waveshape are changed. But with a resistor, unlike the inductor, the energy plane plot expands in all directions as a function of voltage, the plot changes vertically inversely with frequency, and changes shape as a function of the driving waveshape, as seen in the difference between Fig.1-8 (b) and 1-6 (c).

And no matter how many Volt-seconds are applied to the resistor, it will never saturate like the inductor in Fig.1-8 (c).

Noting the striking similarity between the resistor characteristic of Fig. 1-8 (b) and the dash line unsaturated square-loop inductor characteristic of Fig. 1-8 (c), raises an interesting question: If inductance is defined as the slope on the plot of ∫Edt vs. I, then the resistor in (b) is an inductor – it has infinite inductance along B to C and F to G, just like the unsaturated inductor in 1-8 (c).

But if a resistor is defined as a device that does not store energy, only dissipates energy, then the inductor of Fig.1-8 (c) is a resistor!!

Notes on the SMPS Environment

Transformer Definition: A true transformer is a magnetic device with multiple windings whose purpose is *not* to store energy, but to transfer energy instantaneously from input to output(s). In addition, the windings are often electrically insulated to provide high voltage dc isolation between input and output. The turns ratio can be adjusted to obtain optimum relationship between input and output voltages.

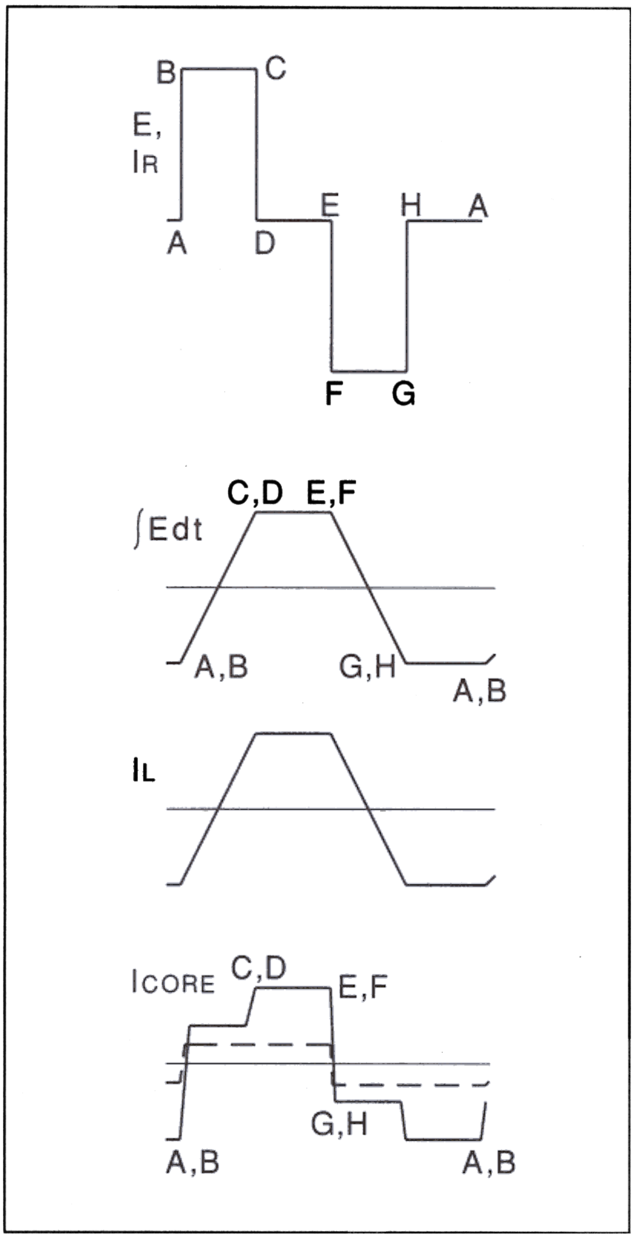


Fig. 1-9 Rectangular Voltage Drive

A practical transformer does store some energy in mutual (magnetizing) inductance and leakage inductances, which degrade circuit performance in several important respects. These inductances are normally considered undesirable parasitics, whose minimization is one of the important goals of transformer design.

Inductor Definition: An inductor is a device whose purpose is to store and release energy. A filter inductor uses this capability to smooth the current through it. A flyback transformer is actually an inductor with multiple windings. It stores energy taken from the input in its mutual inductance during one portion of the switching period, then delivers energy to the output during a subsequent interval.

Since the magnetic core material itself is incapable of storing significant energy, energy storage is accomplished in a non-magnetic gap(s) in series with the core.

Although mutual inductance is an essential element in a flyback transformer, leakage inductances remain undesired parasitic elements.

Core Material Limitations: In dc applications, inductors are thought of as current operated devices. Even the smallest dc voltage will ultimately saturate the magnetic core, unless offset by the IR drop in the winding.

In high frequency SMPS applications, the major core material limitations are saturation and core losses, both of which depend upon flux swing. In these applications, transformer and inductor windings are usually driven with rectangular voltage waveforms derived from low impedance sources. Since the voltage, pulse width, and number of turns are quite accurately known, it is easy to apply Faraday's Law to determine the flux swing and appropriately limit it. In a ferrite core transformer, magnetizing current is difficult to determine accurately. It depends entirely on the core material characteristic which varies widely with temperature and with the flatness of the mating surfaces of the core halves. Fortunately, the magnetizing current in a transformer is small enough to be of less concern than the flux swing.

In an inductor or flyback transformer, the magnetizing current is vitally important, because it represents the energy storage required by the application. In this case, the magnetizing current can be calculated quite accurately using Ampere's Law, because it depends on the very predicable characteristics of the gap in series with the core, and the uncertain core contribution to energy storage is negligible.

Points to Remember:

- Magnetic field equipotentials are surfaces, bounded by the current generating the field.
- All flux lines form complete loops that never begin or end, normal to field equipotentials.
- Flux change cannot occur instantaneously – time is required – an energy change occurs.

- Energy added and removed is quantified by integrating the area between the characteristic and the vertical axis.
- On the energy plane, upward movement in Quadrants I and IV or downward movement in QII and QIII *add* energy to the device. Moving downward in QI and QIV, or upward in QII and QIII returns energy to the circuit.
- The purpose of an inductor is to store energy. In a transformer, energy storage represents an undesired parasitic element.

References

"R-numbered" references are reprinted in the Reference Section at the back of this Manual.

(1) T.G. Wilson, Sr., "Fundamentals of Magnetic Materials," *APEC Tutorial Seminar*, 1987

(R2) "Eddy Current Losses in Transformer Windings and Circuit Wiring," *Unitrode Seminar Manual SEM400*, 1985

Section 2 Magnetic Core Characteristics

Familiarity with the mechanisms underlying magnetic core behavior is essential to (a) optimize the magnetic device design, and (b) properly model its behavior in the circuit application.

The Purpose of the Magnetic Core

The fundamental purpose of any magnetic core is to provide an easy path for flux in order to facilitate flux linkage, or coupling, between two or more magnetic elements. It serves as a "magnetic bus bar" to connect a magnetic source to a magnetic "load".

In a true transformer application, the magnetic source is the primary winding – ampere-turns and volts/turn. The magnetic "load" is the secondary winding (or windings). The flux through the core links the windings to each other. It also enables electrical isolation between windings, and enables adaptation to different voltage levels by adjusting the turns ratio. Energy storage in a transformer core is an undesired parasitic element. With a high permeability core material, energy storage is minimal.

In an inductor, the core provides the flux linkage path between the circuit winding and a non-magnetic gap, physically in series with the core. Virtually all of the energy is stored in the gap. High permeability ferrites, or magnetic metal alloys such as Permalloy, are incapable of storing significant energy. (The integrated area between the nearly vertical high permeability *B-H* characteristic and the vertical axis, representing energy, is minuscule.)

A flyback transformer is actually an inductor with primary and secondary windings and a gap which stores the required energy. Like a simple inductor, the core provides the flux linkage path between the primary and the gap. The core also provides the linkage between the gap and the secondary winding(s) to subsequently deliver the energy to the secondary circuit. Like a transformer, the separate windings also enable electrical isolation between windings, and turns ratio adaptation to different circuit voltages.

Magnetic Core Materials

This paper builds upon Reference (R1), titled "Magnetic Core Properties", taken from an earlier Unitrode seminar and reprinted in the Reference Section at the back of this handbook. It discusses magnetic basics and the process of magnetization in ferromagnetic materials. This topic should be read before proceeding further.

Metal Alloy Tape-Wound Cores

Reference (R1) focuses primarily upon the low-frequency characteristics of metal alloy tape-wound cores. Using alloys such as Permalloy, these cores approach the ideal magnetic material characteristic – square-loop with extremely high permeability (60,000), high saturation flux density (0.9 Tesla = 9000 Gauss) and insignificant energy storage. Unfortunately, resistivity of these metal alloys is quite low. To minimize losses due to induced eddy currents, these cores are built up with very thin tape-wound laminations.

Tape-wound cores are used primarily at 50, 60, and 400 Hz line frequencies. Disappointingly, they are generally unsuitable for transformer applications in SwitchMode Power Supplies. At today's SMPS frequencies (100kHz and up), eddy current losses are too great even with extremely thin 12.5µm (.0005") tape thickness. However, in SMPS filter inductor applications, gapped tape-wound cores are sometimes used when the percent ripple current and associated flux swing is small enough to keep losses at an acceptable level.

Tape-wound cores using the newer, lower loss amorphous metal alloys are used in SMPS applications up to 100-200kHz, especially as magnetic amplifiers.

Powdered Metal Cores

Composite powdered-metal cores, such as powdered iron, Kool Mµ®, and Permalloy powder cores *do* store considerable energy, and are therefore used in inductor and flyback transformer applications. However, energy is *not* stored in the very high permeability magnetic metal portions of the composite, but in the *non-magnetic* regions between the magnetic particles – in the binder that holds the cores together. Essentially, these composite cores store their energy in a non-magnetic gap that is distributed throughout the entire core. These cores are manufactured and categorized by their effective permeability (the permeability of a hypothetical homogeneous core material with the same characteristic as the actual composite). Different effective permeabilities in the range of ≈15 to ≈200 (relative) are achieved by varying particle size and the amount of magnetically inert material in the composite mix.

Composite powdered metal cores are not normally used in true transformer applications because their relatively low permeability results in high magnetizing current and energy storage – undesired in a transformer.

At SMPS frequencies, powdered metal cores are quite lossy. Powdered iron is worst, Kool Mμ is better, Permalloy is best. But in filter inductor or continuous mode flyback applications (where the inductive energy is stored in the non-magnetic regions within the composite core), if the percent ΔI and flux swing are small enough, the losses may be low enough to permit the use of these composite materials.

Rounding of the *B-H* characteristic (which will be discussed later) causes incremental inductance to decrease substantially as the DC operating point is raised. Typically, the inductance may be halved at an operating flux density of 0.4Tesla (4000 Gauss), only half way to saturation.

The much greater saturation flux density B_{SAT} of the powdered metal cores compared to ferrite (0.8T vs. 0.3T) would permit a much smaller inductor as a gapped ferrite for the same application, but at 100 kHz and above, this promise is seldom fulfilled because of the restrictions imposed by losses and rounding.

Ferrite Cores

Ferrites are the most popular core materials used in SMPS applications.

Ferrites are ceramic materials made by sintering a mixture of iron oxide with oxides or carbonates of either manganese and zinc or nickel and zinc. MnZn ferrites are used in applications up to 1 or 2 MHz and include the power ferrite materials used in switching power supplies. NiZn ferrites have lower permeability and much higher resistivity, hence lower losses. They are used from 1 MHz to several hundred MHz.

The permeability of power ferrite materials is in the range of 1500 to 3000 (relative). As shown in the low frequency characteristic of Fig. 2-1, a ferrite core will store a small amount of energy, as shown by the areas between the hysteresis loop and the vertical axis. This undesired magnetizing energy must be subsequently dealt with in a snubber or clamp. Sometimes it can be put to good use in Zero Voltage Transition circuitry. The permeability is high enough to keep the magnetizing current at a generally acceptable level in transformer applications.

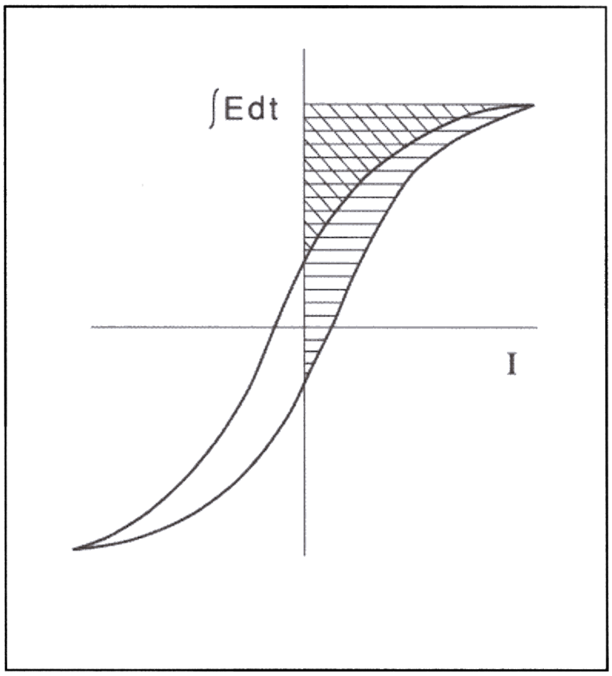


Figure. 2-1 Ferrite Core Characteristic

For inductor and flyback transformer applications, a gap is added in series with the core. This skews the characteristic, and provides the required energy storage capability without the rounding observed in the powdered metal cores.

The reasons for ferrite's popularity in SMPS applications are: lower cost and lower loss than the materials previously discussed. Ferrites are available in a wide variety of core shapes including low-profile and "planar" cores, to facilitate various needs. Twopiece core sets allow the windings to be fabricated separately and subsequently assembled with the core.

The main disadvantage of ferrite is that being a ceramic, the core is less robust than other materials, and may be unacceptable in a high shock military environment.

Saturation flux density in ferrite is much less than with the tape-wound or powdered metal cores: ≈0.3T (3000Gauss) vs. ≈0.8T. This might seem to be a disadvantage, but saturation is not a real limitation at 100kHz or above. In a transformer application, the maximum flux swing is restricted by losses to much less than B_{SAT}. In inductor applications with a small percentage ripple resulting in low core losses, B_{SAT} might become a limiting factor, but the lossier tape-wound or powdered metal cores are usually still at a disadvantage.

Rounding of the B-H Characteristic

Ideal magnetic materials have a square loop characteristic with very high permeability and insignificant stored energy until finally driven into saturation. This is called a "sharp saturation" characteristic. A rounded, or "soft saturation" characteristic exhibits a gradual reduction of incremental permeability until finally the core is completely saturated. Reference (R1) mentions that magnetic "hard spots" and inside corners will cause rounding of the B-H characteristic.

Rounding effects in metal-alloy cores are generally quite trivial. However, in composite powdered metal cores, non-magnetic "gaps" exist between the discrete magnetic particles. Similar non-magnetic inclusions occur among the sintered particles in ferrite cores. These distributed non-magnetic regions cause significant rounding of the *B-H* characteristic. They also result in storing energy within the core. The particulate structure has two main effects:

First, the distributed reluctance of these tiny "gaps" causes the flux and the flux change to be distributed across the entire core, rather than as a discrete flux change boundary moving from inside to outside as depicted in (R1) for ungapped idealized metal alloy cores.

Second, at low flux densities, flux tends to concentrate in the "easiest" paths (lowest reluctance) where the magnetic particles are in close proximity. As the flux density increases, these easy path areas are the first to saturate. Those portions of the magnetic particles that saturate first become non-magnetic, making these paths less "easy". Incremental flux increase shifts to adjacent paths where the magnetic material has not yet saturated and where the gap is somewhat wider. This process continues, effectively widening the incremental distributed gap as the flux increases. The incremental permeability (and inductance) is progressively reduced, as observed in the rounding of the *B-H* characteristic.

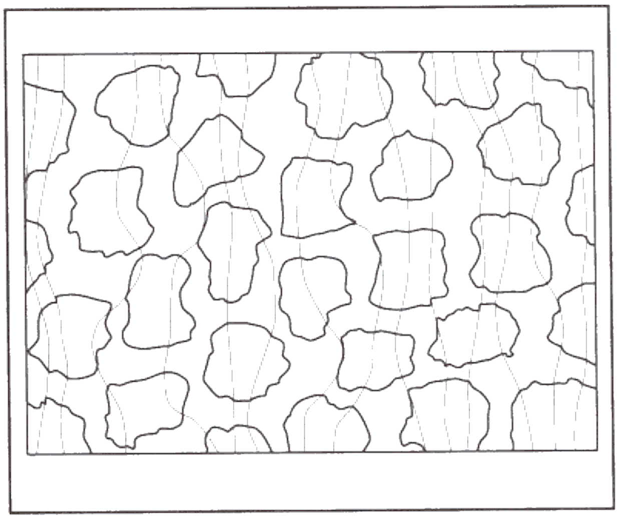


Figure. 2-2 -- Easy Flux Path Between Particles

In powdered metal cores, this non-linear inductance characteristic is unavoidable, except by restricting the maximum flux density to a small fraction of B_{SAT}. In some filter inductor applications, the rounding effect, akin to a "swinging choke", might actually be desirable.

In a ferrite core, the rounding effect is, if anything, beneficial. In a transformer application, normal operation with flux density limited by core losses, the rounding is not encountered, and even if it is, the result is a small increase in magnetizing current. In a situation where the flux "walks" toward saturation due to a volt-second imbalance on the transformer primary, the soft saturation characteristic provides a gradual magnetizing current increase to facilitate control of this problem.

When a discrete gap is added to the ferrite core for energy storage in a filter inductor application, the rounding of the ferrite characteristic disappears – swamped by the linear high reluctance of the gap. The inductance characteristic becomes quite linear until saturation is reached. If a non-linear inductance characteristic is desired, it can be accomplished with a tapered or stepped gap.

Core Limitations in SMPS Applications

In high frequency SwitchMode power supplies, magnetic core characteristics usually impose different limitations in transformer, filter inductor, and flyback transformer applications.

A true transformer is commonly used in buck-derived circuits such as the forward converter, full bridge, half bridge, etc. Ideally, a transformer stores no energy, but transfers energy immediately from input to output. In a practical transformer, undesired stored energy does occur in parasitic leakage inductances (outside the core), and magnetizing inductance (within the core). Magnetizing inductance is minimized by using a gapless, high permeability core material.

At low frequencies, core saturation is usually the most important limitation. But at SMPS frequencies, usually 100kHz or greater, *core loss* becomes the most important limitation in transformer applications. Powdered metal cores are effectively ruled out because of high losses and because of low permeability.

Tape-wound metal alloy cores have considerably higher core losses than ferrite cores. Tape-wound cores have higher B_{SAT} than ferrite, but this is irrelevant because core loss severely restricts the flux swing. Tape-wound cores are considered for SMPS transformer applications only if their greater ruggedness is needed.

A **filter inductor** must store energy during one portion of each switching period and return this energy to the circuit during another portion of the period, thus smoothing the current flow. The required energy must be stored in a non-magnetic gap – distributed in the case of a powdered metal core, and a discrete gap in series with a ferrite core or tape-wound metal alloy core.

If the switching frequency and the percentage of current ripple (which determines the flux swing) are both low enough, core losses will be low, and the inductor core may be limited by saturation. In this situation, powdered metal cores or gapped tape-wound cores may not only be feasible, they *may* outperform gapped ferrite cores because of their higher B_{SAT}. But with higher frequency and/or larger percent ripple current, core losses will dominate, and ferrite cores will outperform the others.

In situations where powdered metal cores may be advantageous, bear in mind that the inductance may diminish an unacceptable amount at higher current levels due to the rounding effect discussed earlier.

Flyback transformers are really inductors with multiple windings. There are some unique problems associated with the windings, but the core does not care how many windings exist – the core is aware only of the total ampere-turns and the volts/turn. When operated in the continuous current mode, with small ΔI and at low enough frequency, the same considerations apply as for the simple inductor. In the discontinuous mode, the current swing (and flux swing) become very large, the core loss limitation applies, and gapped ferrite cores provide the best performance.

Transformer and inductor design is covered in detail in other sections of this manual.

Core Saturation

At SMPS switching frequencies, core saturation is almost never a limitation in transformer applications, although it often is in filter inductors or continuous mode flyback transformers.

Flux Walking: Transformers operated in push-pull circuits *do* have a potential problem with core saturation.

A positive pulse applied to a winding causes a positive flux change proportional to the pulse volt-seconds. In order to maintain a stable operating point on the *B-H* characteristic, the core must be "reset" by subsequently applying the exact same number of negative volt-seconds.

In a single-ended application, such as a forward converter, the core "resets itself" by an inductive voltage reversal which self-terminates when the magnetizing current returns to zero. In a push-pull application, the core is reset by the circuit, which applies sequential positive and negative pulses to the windings. With the slightest asymmetry – inequality of either voltage or time – the positive and negative volt-seconds do not completely cancel. As a result, the flux never quite returns to its starting point, and over a period of many cycles at the switching frequency, the flux density "walks" into saturation. This problem is not a core limitation – any core would eventually reach saturation. This is a circuit problem, to which there are several circuit solutions which are beyond the scope of this paper.

Core Loss

Core loss is the most important core limitation in most SMPS applications. For acceptable losses, flux density swing ΔB must be restricted to much less than B_{SAT}. This prevents the core from being utilized to its full capability.

At low frequencies, core loss is almost entirely hysteresis loss. For today's power ferrites, eddy current loss overtakes hysteresis loss at 200-300kHz. In metal alloy cores, eddy current loss dominates above a few hundred Hertz.

Core manufacturers usually provide curves such as Fig. 2-3 showing core loss as a function of flux swing and frequency, combining hysteresis and eddy current losses. Core loss is usually expressed in mW/cm^3, sometimes in kW/m^3 (actually equal: 1 mW/cm^3 = 1 kW/m^3), sometimes in Watts/pound (horrors!!)

In these Core Loss vs. Flux Density curves, the horizontal axis labeled "Flux Density" usually represents *peak* flux density, with symmetrical sinusoidal excitation. In SMPS applications, *peak-to-peak* flux swing, ΔB, is calculated from Faraday's Law, where $\int Edt$ = applied Volt-seconds, N = turns, and A_e = core cross-section area:

$$\Delta B = \frac{1}{NA_e}\int Edt$$

The total flux swing, ΔB, is twice the peak flux swing referred to in the core loss curves as "Flux Density". Therefore, use $\Delta B/2$ to enter the core loss curves.

Hysteresis Loss

The hysteresis loops shown in the core material data sheets represent the core overdriven by a sinusoidal waveform from + to – saturation. In an SMPS application, the core is usually driven by a much smaller rectangular waveform with ΔB limited by core losses to a minor hysteresis loop as shown in Fig. 2-4.

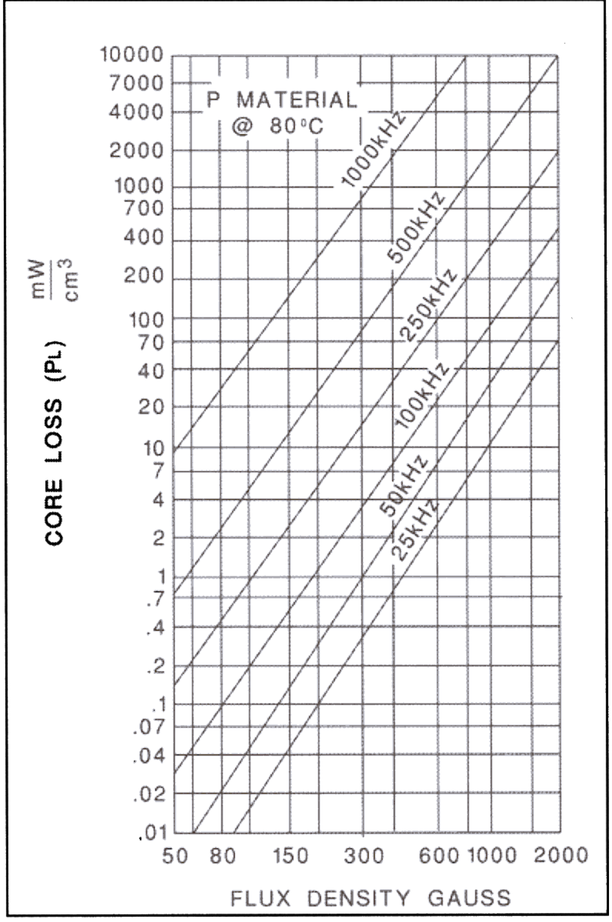


Figure. 2-3 Core Loss -- "P" Material

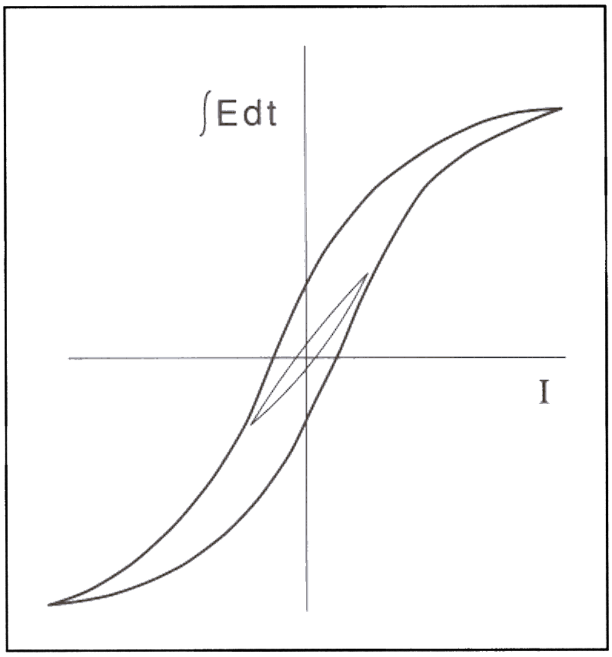


Figure. 2-4 – Minor Hysteresis Loop

The hysteresis loop area represents energy loss. Power loss depends on how many times per second the hysteresis loop is traversed. Thus, hysteresis loss varies directly with frequency.

Hysteresis loss varies with flux density swing (ΔB) to some power, depending on how much the minor hysteresis loop expands horizontally (ΔH) as well as vertically (ΔB). For most ferrites, hysteresis loss varies with ΔB^n, with $n \approx 2.5 - 3.0$.

The hysteresis loop changes shape somewhat with waveshape, current or voltage drive, and temperature. This variability, together with the steep slope of the high inductance characteristic makes it impossible to predict the magnetizing current with any degree of accuracy (and eddy currents make the problem worse).

Fortunately, the only important concern about the hysteresis loop in SMPS applications is the core loss it represents. The shape does not matter – the core loss curves provide the necessary information. In transformer applications, all we really need to know is that the magnetizing current is acceptably low (unless I_m is depended upon for some circuit function, which is risky). In filter inductor and flyback transformer applications, the hysteresis loop of the core material is totally swamped by the lossless and predictable high reluctance of the series gap, making I_m easily predictable.

Eddy Current Loss

At the high frequencies usually involved in SMPS applications, it is incorrect to think of eddy current losses in the core as being frequency dependent. Core eddy current loss is a function of the volts per turn applied to the windings, and the duty cycle. It can be modeled by placing a resistor across one of the windings.

For example, a square wave of 5 Volts/turn, as shown in Fig. 2-5, applied to the primary, will result in the same eddy current loss *regardless of frequency.* On the other hand, if the waveform is changed to 10 Volts peak at 50% duty cycle (same average value), the peak loss quadruples (proportional to V_p^2) while the average loss doubles (the duty cycle is halved), whether the frequency is constant or not. The resistor model demonstrates this behavior.

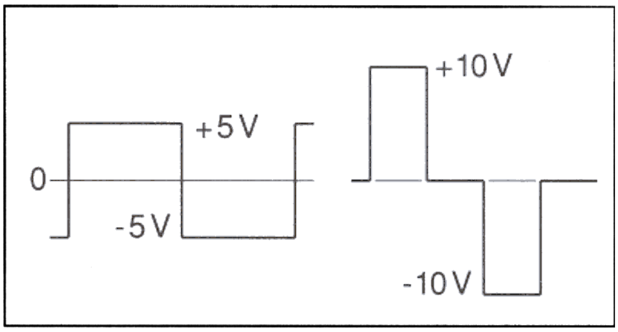


Figure. 2-5 Waveform Comparison

What causes eddy currents: Virtually all of the flux induced by the primary winding is contained within the core. The core itself is a single turn secondary linked to all of the windings. A voltage is induced around the core periphery equal to the volts/turn applied to the windings. The core material has a finite resistivity, which translates into a resistance value around the core periphery. The voltage induced around the core forces a current – the eddy current – to flow through this resistance. The result is I^2R loss. The eddy current is reflected into the primary according to the ratio of the primary turns to the single turn "core secondary". In the primary, it is considered part of the magnetizing current, although it is pure loss, and in fact absorbs some of the stored energy in the core.

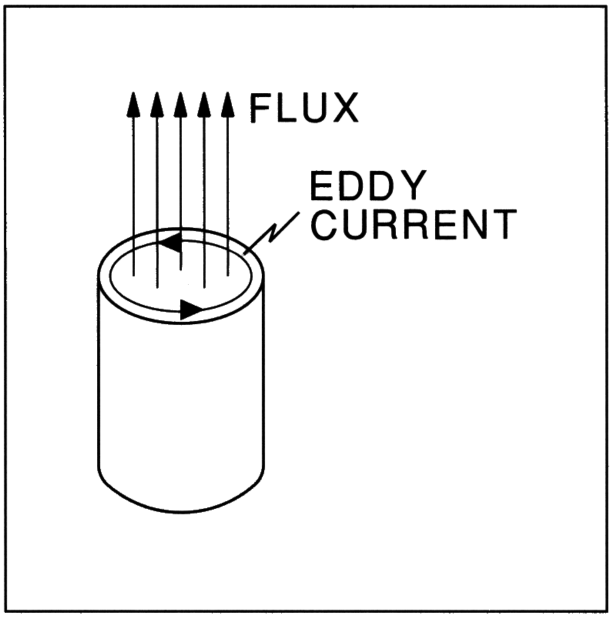


Fig. 2-6 Eddy Current in a Ferrite Core

The effective eddy current resistance can be calculated from the core resistivity and dimensions, and projected as an equivalent resistor across the primary winding according to the turns ratio squared. Hence the validity of the resistor model mentioned earlier. Placing the equivalent eddy current resistance across the primary, it is obvious that if the applied voltage changes, the eddy current instantaneously changes proportionally. When the voltage pulse ends and the voltage drops to zero, the eddy current becomes zero. The magnetizing current then reverts to the low frequency hysteresis loop value.

Resistivity of **power ferrite** materials intended for SMPS applications ranges from 200 to 2000 Ω-cm. The resulting eddy currents and associated losses are barely noticeable compared to hysteresis effects at 100 kHz, but become dominant in the range of 250-600 kHz, depending on the specific material.

Resistivity of **tape-wound metal alloy** cores is only 50 to 150 *micro*-Ω-cm. If the core was solid metal, it would be a shorted turn. High current would circulate on the core surface, and the magnetic field could not penetrate the core.

The solution to this problem is to break the core up into electrically insulated laminations. Figure 2-7 shows the detail of one such lamination. If a metal alloy core 1.2 cm thick is divided into 1000 laminations, each .0012 cm thick, then each lamination contains only 1/1000 of the total flux. Therefore the voltage induced around the periphery of each lamination is 1/1000 of the volts/turn in the windings. The resistance around the periphery of the lamination (the eddy current path) is ≈500 times the resistance of the path around the entire core if it were solid. ($R=\rho\ell/A$; path length ℓ around the lamination is ≈ ½ the path length around the entire core periphery; area A is ≈ 1/1000 of the effective area around the periphery of the solid core).

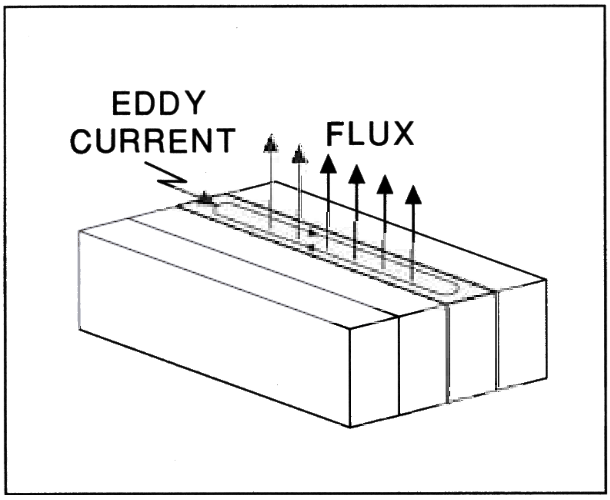


Figure. 2-7 Lamination Detail

Thus the effective eddy current resistance is roughly equivalent to a solid core with resistivity 500,000 times greater. This is beginning to approach a solid ferrite core.

Skin Effect

The example of the solid metal alloy core discussed earlier raises the concern that the magnetic field may not be able to penetrate the core sufficiently, confining the flux to the core surface. A calculation of the penetration depth resolves this issue.

Penetration depth (skin depth) is defined as the distance from the surface to where the current density is 1/e times the surface current density:

$$D_{PEN} = \sqrt{\frac{\rho}{\pi \mu_0 \mu_r f}} \quad \text{meters}$$

In a **Permalloy** tape-wound core, resistivity, $\rho = 55 \cdot 10^{-4}$ Ω-m, and $\mu r = 30{,}000$:

$$D_{PEN} = 0.22 / \sqrt{f} \quad \text{cm}$$

At 100kHz in Permalloy, D_{PEN} = .0007 cm. Thus, even with .0005" (.00125cm, or 12.5μm) tape thickness, there is some concentration of current at the tape surfaces, slightly worsening the eddy current losses.

In power **ferrite** (Magnetics type K), with $\rho = 20$ Ω-m and $\mu_r = 1500$:

$$D_{PEN} = 5800 / \sqrt{f} \quad \text{cm} \qquad \text{(ferrite)}$$

At 100kHz, in ferrite Type K, D_{PEN} = 18cm. Penetration depth in ferrite is thus much greater than core thickness, and skin effect can be ignored.

Modeling the Magnetic Core

The subject of modeling the complete magnetic device is discussed in Reference R3. The computer model might include:

- An ideal transformer with appropriate turns ratios for the windings
- leakage inductances properly located and quantified
- parasitic capacitances
- A model of the magnetic core, reflected into the primary winding or a dedicated winding.

In this Section, where the core is under discussion, it is appropriate to consider some aspects of modeling the core.

Based on the preceding discussion of the mechanisms underlying magnetic core behavior, there are several distinct elements that might be included in the core model:

- Inductance, magnetizing current, linear region, rounding, saturation – ungapped and gapped.
- Hysteresis loop centered on the inductance characteristic
- Eddy current loss – proportional to dB/dt

Physical or Empirical ???

Should the core model be based on physical mechanisms or on empirical data? Core eddy current can easily be modeled based on the physical process, simply by placing a resistor across the primary, or across a normalized one-turn winding dedicated to hold the core model.

But for the other core characteristics, discussed earlier in this paper, the underlying physical processes are much too complex to serve as the basis for the core model. An empirical model is the only reasonable choice. However, familiarity with the physical processes *is* helpful in developing an empirical model.

There are many types of evaluations used to understand, improve, or validate various aspects of SMPS operation. Rather than use the same complex "one size fits all" model for all purposes, it is far better to use simpler models tailored to the needs of each specific situation. The evaluations will be faster, and the results are easier to understand. Include only the characteristics that are necessary, and for these, use the simplest approximation that will serve the purpose. Only when a really accurate representation is required, use a mathematical expression, curve fitted to the data sheet representation.

Core characteristics that are most important in SMPS applications are (1) losses; (2) saturation. Sometimes, in a transformer, a crude representation of magnetizing current is necessary. Replicating the curvature of the characteristic is not important except in filter inductor applications using powdered metal cores operated close to saturation, which results in increased ripple current at full load. Other than that situation, it will usually suffice to have two straight lines, one representing normal inductance, intersecting at BSAT with a second straight line representing saturated inductance. If it is certain that the core will not be driven into saturation in the evaluation being performed, the saturated inductance line may be omitted.

An adequate model for different core materials in each of the following applications might include:

- **Filter inductor with gapped ferrite core:** Until the core saturates, the gap is the only important element determining the inductance value – the effect of the ferrite core is negligible. For example: With a 0.3 cm gap and a 10 cm path length in the ferrite (μ_r = 3000), the core reduces inductance by 1%. Since current limiting prevents the inductor from being driven into saturation, a simple, linear inductor model, representing the gap, is generally adequate. In computer circuit modeling, the standard L model can be used, with current specified as initial condition.
- **Filter inductor with gapped, laminated metal-alloy core:** The same considerations as the gapped ferrite core will apply. A shunt resistor to model core eddy-current losses may be desirable with this lossier material.
- **Filter inductor with powdered metal core** – MPP, Powdered iron, Kool Mμ®: If flux density is limited by core loss to below the region of curvature, the simple linear inductance model may suffice. If operated at higher flux densities where inductance non-linearity becomes an important consideration, a customized computer model should be used, curve-fitted to the actual inductance characteristic.

- **Saturable reactor (magamp) core:** Mag-amps require high permeability square-loop core materials – metal alloy laminated cores or square-loop ferrites. It is essential to include the hysteresis characteristic in a magamp core model. In metal alloy cores, hysteresis *loss* is negligible compared to eddy current loss at SMPS frequencies. Nevertheless, if the square loop hysteresis characteristic is not included in the model, the flux will be unable to maintain its position on the vertical axis after reset is accomplished and before the next power pulse, and will drift toward zero.

  A magamp core model requires hysteresis, eddy current, and saturation.
- **Transformer with ungapped metal alloy core**: A straight line representing the very high unsaturated inductance is probably acceptable. An intersecting straight line representing the saturated inductance could be included if necessary for the intended evaluation. Eddy current loss resistance is very significant. Hysteresis can be omitted.
- **Transformer with ungapped ferrite core**: As above, straight line approximations can be used to simulate inductance and saturation. Losses are primarily hysteretic at 100-200 kHz. Eddy current losses become significant at higher frequencies.

Figure 2-8 shows how the elements representing the core can be applied to a transformer model. Eddy current resistance is modeled by resistor R_E. All of the other core characteristics are included in the symbol labeled "CORE" – inductance, rounding, saturation, and hysteresis effects, but only to the degree and accuracy relevant to the intended circuit evaluation.

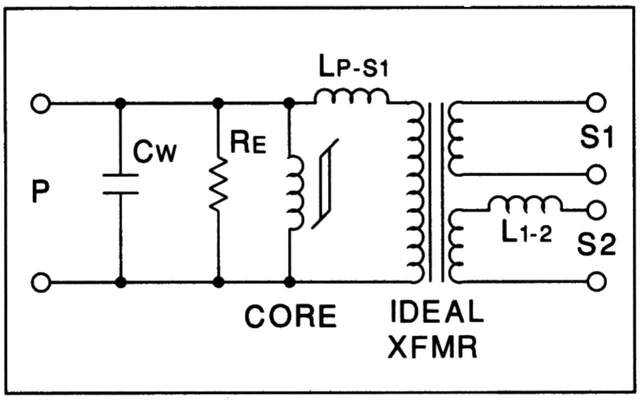


Fig. 2-8 Transformer Model

Points to Remember

- Hysteresis loss is directly proportional to frequency, and to the n^{th} power of the flux density swing. n is in the range of 2.5 to 3.0 for most power ferrites.
- Eddy current in the core is proportional to $d\phi/dt$ (equal to volts/turn) not frequency per se. Eddy current *loss* (I^2R) is proportional to volts/turn *squared.* When the voltage transitions to zero at the end of each pulse, the eddy current becomes zero.
- At DC and low frequency, magnetic devices are current-controlled. At the SMPS switching frequency, they are voltage driven.
- Core limitations – saturation and losses – put restrictions on flux and flux swing, which translate into volt-second limitations on the applied voltage waveforms.

References

"R-numbered" references are reprinted in the Reference Section at the back of this Manual.

(R1) "Magnetic Core Properties," originally titled *"An Electrical Circuit Model for Magnetic Cores,"* Unitrode Seminar Manual SEM1000, 1995

(R3) *"Deriving the Equivalent Electrical Circuit from the Magnetic Device Physical Properties"*, Unitrode Seminar Manual SEM1000, 1995 and SEM1100, 1996

Section 3 Windings

Understanding the rules governing magnetic field behavior is fundamentally important in designing and optimizing magnetic devices used in high frequency switching power supply applications. Paralleled windings can easily fail in their intended purpose, eddy current losses and leakage inductances can easily be excessive. These are some of the problems that are addressed in this Section.

Even if you never participate in transformer or inductor design, these magnetic principles apply in optimizing circuit layout and wiring practices, and minimizing EMI..

Reference paper (R2): "Eddy Current Losses in Transformer Windings and Circuit Wiring," included in this Manual, is a useful supplement.

Conservation of Energy

Like water running downhill, electrical current always takes the easiest path available. The path taken at dc and low frequencies can be quite different from the path taken by the high frequency current components.

The basic rule governing the current path: *Current flows in the path(s) that result in the lowest expenditure of energy*. At low frequency, this is accomplished by minimizing I^2R losses. At high frequency, current flows in the path(s) that minimize inductive energy – energy transfer to and from the magnetic field generated by the current flow. Energy conservation causes high frequency current to flow near the surface of a thick conductor, and only certain surfaces, even though this may results in much higher I^2R losses. If there are several available paths, HF current will take the path(s) that minimize inductance. This may have undesirable side effects as shown in one of the examples below.

Examples are given later which demonstrate how to manipulate the field and current path to advantage.

Skin Effect

The circuit of Figure 3-1 shows an inductor in series with an L-R transmission line. What happens when a dc current is put through this circuit? What

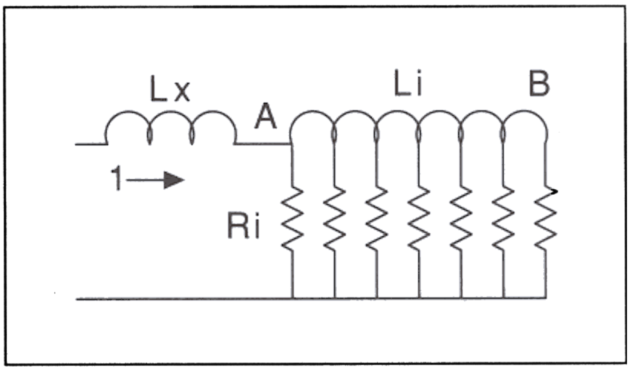


Figure 3-1 Skin Effect Model

happens when a high frequency ac current is put through?

Figure 3-1 happens to be a high-frequency model of a single wire. A represents the surface of the wire, B is the center. Lx represents the inductance per unit length *external* to the wire (what would be the measured inductance of the wire). Li is inductance distributed within the wire, from the surface to the center. (Copper is non-magnetic, just like air, and stores magnetic energy in the same way.) Ri is the distributed longitudinal resistance from the surface to the center. Collectively, Ri is the dc resistance of the wire. All of the above values are per unit length of wire.

At dc or low frequency ac, energy transfer to inductance Li, over time, is trivial compared to energy dissipated in the resistance. The current distributes itself uniformly through the wire from the surface to the center, to minimize the $I^2R{\cdot}t$ loss. But at high frequency, over the short time spans involved, $I^2R{\cdot}t$ loss is less than the energy transfer to and from Li. Current flow then concentrates near the surface, even though the net resistance is much greater, in order to minimize energy transfer to Li. If we look at this strictly from a circuit point of view, at high frequency, the impedance of Li near the surface blocks the current from flowing in the center of the wire.

Penetration depth (or skin depth), *Dpen*, is defined as the distance from the conductor surface to where the current density (and the field, which terminates on the current flow) is 1/e times the surface current density:

$$D_{PEN} = \sqrt{\frac{\rho}{\pi\mu_0\mu_r f}} \quad \text{meters}$$

In **copper** at 100°C, resistivity, $\rho = 2.3 \cdot 10^{-8}$ Ω-m, and $\mu_r = 1$:

$$D_{PEN} = {}^{7.6}\!/\!\sqrt{f} \quad \text{cm} \qquad \text{(copper)}$$

At 100 kHz in copper, $D_{PEN} = .024$ cm.

Proximity Effect

When two conductors, thicker than D_{PEN}, are in proximity and carry opposing currents, the high frequency current components spread across the surfaces facing each other in order to minimize magnetic field energy transfer (minimizing inductance) This high frequency conduction pattern is shown in reference (R2) Figs 5, 6, and 7. (R2) Fig. 5 is reproduced below as Fig. 3-2.)

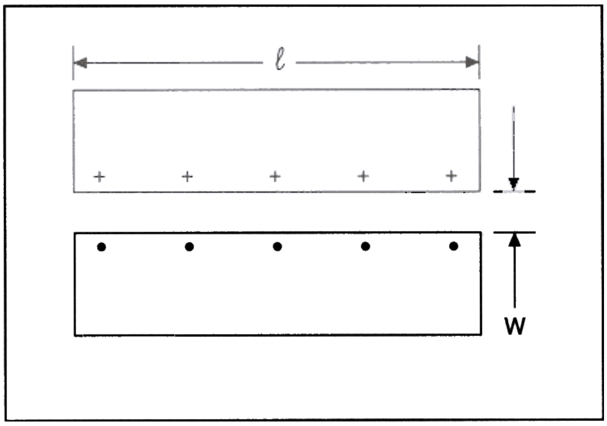


Fig. 3-2 Proximity Effect

In all three configurations, current does not flow on any other surfaces because that would increase the volume of the magnetic field and require greater energy transfer. Inductance is thus minimized, but ac resistance is made higher, especially in (R2) Figs 6 and 7. Note how in Fig. 3-2, the preferred configuration, the current distributes itself fairly uniformly across the two opposing surfaces. This results in significantly less stored energy than R2 Fig. 6, even though the length and volume of the field in Fig. 3-2 is 5 times greater. This is because spreading the current over 5 times longer distance reduces the field *intensity*, H, by 5 times. Energy density, proportional to H^2, is 25 times weaker.

$$(W/cm^3 = \tfrac{1}{2} BH = \tfrac{1}{2} \mu H^2)$$

Therefore, total energy (volume times energy) and inductance, are 5 times less in Fig. 3-2 above than with the more concentrated field in (R2) Fig. 6.

An important principle is demonstrated here: If the field (and the current that produces it) is given the opportunity to spread out, it will do so in order to minimize energy transfer. The stored energy (and inductance) between the conductors varies inversely with the length of the field.

Visualize the magnetic field equipotential surfaces stretched across the space between the two conductors, terminating on the current flow at each surface. Visualize the flux lines, all passing horizontally between the two conductors, normal to the equipotential surfaces. The flux return paths encircle the conductors in wide loops spread out over a distance – the field here is very weak.

Examples

In the examples of winding structures given below:

- Each + represents 1 Ampere into the page
- Each • represents 1 Ampere out of the page
- Fine lines connecting + and • represent edge view of the field equipotential surfaces.
- Conductor size is much greater than penetration (skin) depth.

Simple Transformer Windings:

If the two flat conductors of Fig. 3-2 are placed within a transformer core, the only change in the field pattern is that the fringing field at the conductor ends is reduced. The conductors are now called "windings", and the inductance representing the energy stored between the conductors is called "leakage inductance". In Figure 3-3, one of the flat strips is replaced by 4 wires. This could be a 4-turn winding carrying 3A, opposed by a single turn secondary carrying 12A. *Or,* the four primary wires could be in parallel, giving a 1-turn primary carrying 12A. In either case, the field pattern spreads itself across the entire window, and the same minimized energy is

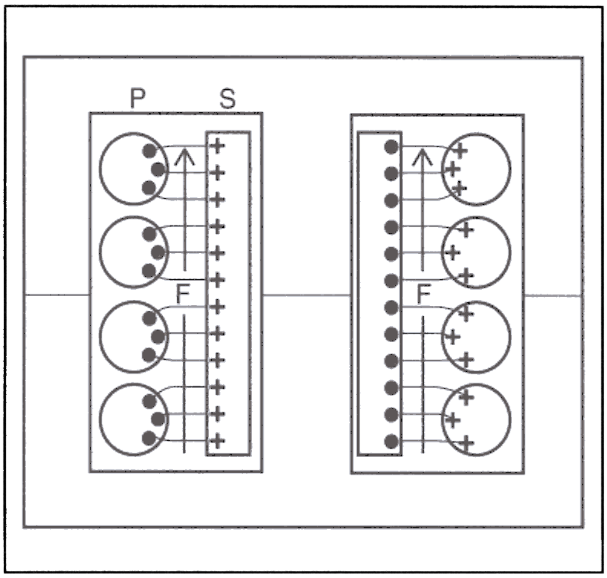


Figure 3-3 Single Layer Windings

stored between the windings. The conductors are thicker than D_{PEN}, so the high frequency currents flow near the surfaces in closest proximity, thus terminating the field.

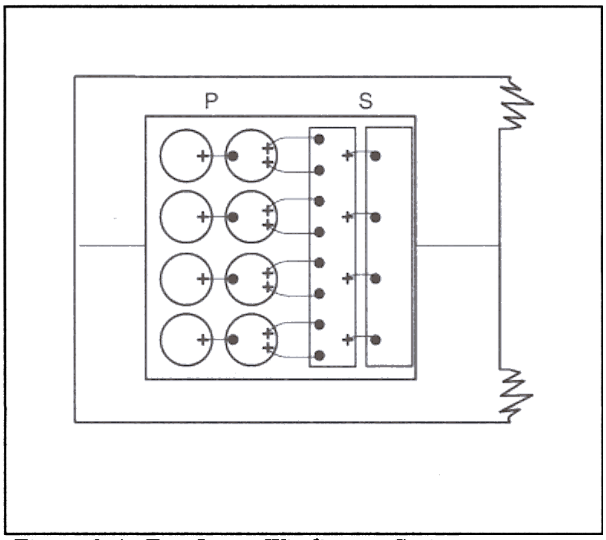


Figure 3-4 Two Layer Windings -- Series

Multiple Layer Windings:

In Figure 3-4, an 8-turn primary carrying 1A is opposed by a 2-turn secondary carrying 4A. The 8 turns of wire, sized for the required rms current, cannot fit into the window breadth, so it is configured in two layers. As expected, there is an 8 Ampere-turn field stretched across the entire window. But since the conductor thickness is much greater than the skin depth, the field cannot penetrate the conductor, and current flow is confined to near the conductor surface.

A strange thing happens -- since the field cannot penetrate the conductor, the entire 8 Ampere field terminates at this inner surface of the inner layer. This requires a total of 8 Ampere-turns at this surface -- 2A per wire -- since the field can be terminated only by current flow. Inside the outer layer, there is a 4A field, from the 4 A-t flowing in the outer layer. This field must terminate on the outside of the inner layer, because it cannot penetrate. This requires 4 A-t in the *opposite direction* of the current in the wire!!

Thus the inner layer has 8 A-t on its inner surface, and 4 A-t in the opposite direction on its outer surface. Each inner wire has 2A on its inner surface and 1A in opposite direction on its outer surface. The net current remains 1A in all series wires in both layers. But since loss is proportional to I^2, the loss in the inner layer is $1^2 + 2^2 = 5$ times larger than the loss in the outer layer, where only the net 1A flows on its inner surface!!

Not only is the I^2R loss larger because the current is confined to the surface, it also increases *exponentially* as the number of layers increases. This is because the field intensity increases progressively toward the inside of the winding. Since the field cannot penetrate the conductors, surface currents must also increase progressively in the inner layers. For example, if there were 6 wire layers, all wires in series carrying 1A, then each wire in the inner layer will have 6A flowing on its inner surface (facing the secondary winding) and 5A in the opposite direction on its outer surface. The loss in the inner layer is $6^2 + 5^2 = 61$ times larger than in the outer layer which has only the net 1A flowing on its inner surface!!

If the wire diameter is reduced, approaching the penetration depth, the + and – currents on the inner and outer surfaces of each wire start to merge, partially canceling. The field partially penetrates through the conductor. When the wire diameter is much less than the penetration depth, the field penetrates completely, the opposing currents at the surfaces completely merge and cancel, and the 1A current flow is distributed throughout each wire.

Calculation of the I^2R loss when the conductor size (layer thickness) is similar to the penetration depth is very complex. A method of calculating the ac resistance was published by Dowell[1], and is discussed extensively in Reference (R2). Figure 3-5 (from R2, Fig. 15), based on Dowell's work, shows the ratio of R_{AC}/R_{DC} vs. layer thickness/D_{PEN} and the number of layers in each winding section. Read (R2) for a detailed discussion.

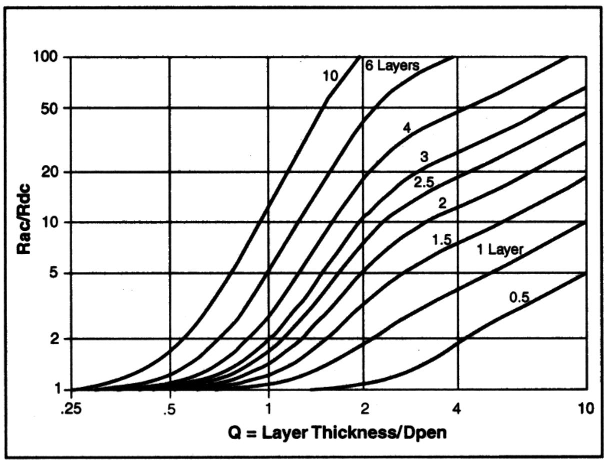


Figure. 3-5 Eddy Current Losses -- R_{AC}/R_{DC}

The curves of Fig. 3-5 graphically show the high ac resistance that results when the layer thickness equals or exceeds the penetration depth, especially with multiple layers. With the large ac currents in a transformer, R_{AC}/R_{DC} of 1.5 is generally considered optimum. A lower R_{AC}/R_{DC} ratio requires finer wire, and the wire insulation and voids between wires reduce the amount of copper, resulting in higher dc losses. In a filter inductor with small ac ripple current component, a much larger R_{AC}/R_{DC} can be tolerated.

Although the curves of Fig. 3-5 are quite useful, keep in mind that an accurate solution requires harmonic (Fourier) analysis of the current waveform. Loss must then be calculated independently for each harmonic, since D_{PEN} differs for each harmonic frequency, and these losses added to obtain the total loss.

Alternative methods of calculating eddy current losses include:

1. Calculate based on the fundamental only, ignore the harmonics and add 50% to the calculated loss.

2. Carsten[2] has applied Dowell's sine wave solution to a variety of non-sinusoidal waveforms encountered in SMPS applications, providing curves that include harmonic effects.

3. Computerize Dowell's work, in order to apply it to any non-sinusoidal waveshape. O'Meara[3] can be helpful.

4. A computer program "PROXY" (proximity effect analysis) is available from KO Systems[4].

Paralleled Layers:

The transformer of Fig. 3-6 is the same as Fig. 3-4 with the winding layers reconfigured in parallel, resulting in a 4-turn, 2 Amp primary, and a 1-turn, 8A secondary. The intention is to have 1A in each primary wire and 4A in each secondary strip, with the same field pattern as in Fig. 3-4, but it doesn't happen that way.

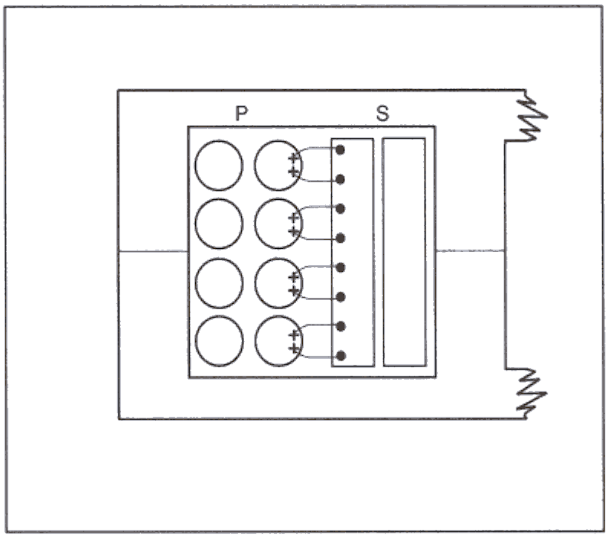


Figure 3-6 Paralleled Two-Layer Windings

Whenever windings are paralleled, alternative current paths are provided. In Fig. 3-6, resistance in each of the paralleled windings causes the current to divide nearly equally at dc and low frequencies. But at high frequency, stored energy becomes more important than $I^2R{\bullet}t$. *All of the high frequency current components will flow on the inside surfaces of the inner layers directly facing each other.* The high frequency current in the outer layers is *zero*. Any current in the outer layers contributes to an additional field, between inner and outer layers, requiring additional energy. With series-connected layers the current has no alternative – it *must* flow in all layers, resulting in additional energy stored between layers as shown in Fig. 3-4. When there is an alternative, as with paralleled layers, the high frequency current will flow so as to minimize the stored energy.

The leakage inductance between the windings in Fig 3-6 is slightly smaller than it would have been if the current divided equally – a small benefit. But only a tiny fraction of the available copper is utilized, making I^2R loss prohibitively large.

Another example: A low voltage, high current secondary might use a single turn of copper strap, but the thickness required to carry the rms current is 5 times the skin depth. It might seem logical to parallel 5 thin strips, each one skin depth in thickness. Result: All the HF current will flow in the one strip closest to the primary. If ac current were to flow in the other strips further from the primary, I^2R loss would be less, but more stored energy is required because the field is bigger (increased separation).

Rule: If you provide alternative current paths, be sure you know what the rules are.

Window Shape–Maximize Breadth

The shape of the winding window has a great impact on the eddy current problem. Modern cores intended for high frequency SMPS applications have a window shape with a winding breadth (width) several times greater than its height. For the same number of turns, the number of layers required is thereby minimized. As shown in Figure 3-7, the window has twice the breadth as the core in Fig. 3-4, so that only one layer is required. This results in a very significant reduction in eddy current losses, as can be seen from Fig 3-5.

Another major benefit of the wider window is that the stored energy (leakage inductance) is minimized. With the 8 turns at 1A of Fig 3-4 fit into a single layer, (and the opposing 2-turn strap also in a single layer), the total magnetic force, ι, remains 8 Ampere-turns. However, the field intensity *H,* stretched over twice the distance, is half as much. Flux density *B* is also halved ($B = \mu H$), therefore energy density $\frac{1}{2}BH$ is one-fourth as much. However, the volume of the field is increased, perhaps doubling, therefore total energy–and leakage inductance–is halved.

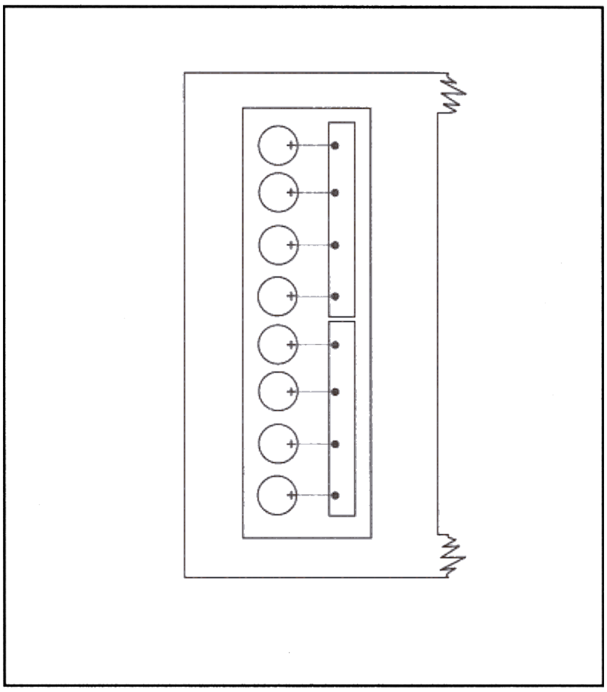


Figure. 3-7 Wide Window Breadth

The penalty for stretching out the winding is increased capacitance between windings.

Interleaving:

If the stretched out winding of Fig. 3-7 were folded in half, it could then fit into the original window, as shown in Figure 3-8. This "interleaved" winding has the same low eddy current loss, low field intensity, and low inductance as the winding of Fig. 3-7.

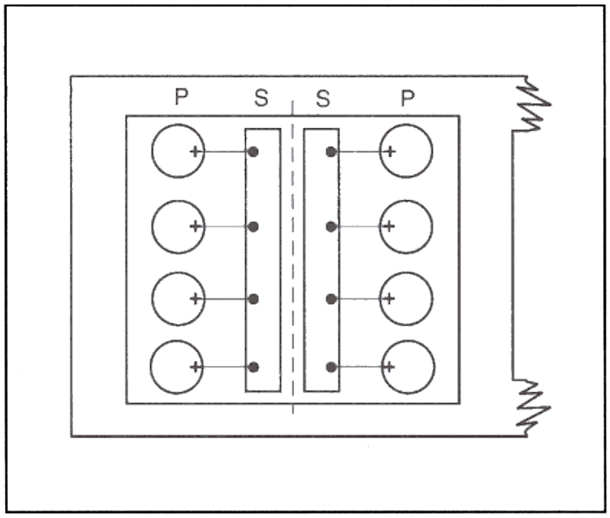


Figure. 3-8 Interleaved Windings

Winding Sections:

Although there are two primary winding layers and two secondary layers in Fig. 3-8, the interleaved windings actually divide into two *winding sections,* indicated by the dashed line between the secondary layers. The boundary between winding sections is defined as the point where the total magnetic field goes through zero. Within each winding section of Fig. 3-8, there is a 4 Ampere-turn field introduced by the primary layer and canceled by the opposing secondary layer. At the dashed line between the two secondary layers, the total field is zero. Thus, in Fig. 3-8, there is one primary layer and one secondary layer in each of two winding sections. Further detail, including the concept of half layers seen in Fig. 3-6, is given in (R2).

First level interleaving (2 winding sections) is very beneficial in terms of eddy current loss and leakage inductance. EMI is minimized because the fields in each winding section are oppose each other externally. The penalty for interleaving is increased primary to secondary capacitance.

With further levels of interleaving, gains become marginal and the interwinding capacitance penalty gets worse.

A 3 winding section structure (P-S-P-S) is often executed incorrectly. For proper balance, the fields should be equal in each of the 3 winding sections. This requires the two interior winding portions have twice the Ampere-turns of the two outside winding portions: (P-SS-PP-S).

When does Paralleling Succeed:

Paralleling succeeds when the equal division of high frequency current among the parallel paths results in the least stored energy. Paralleling fails when unequal division of current results in the least stored energy. High frequency current will always take the path that results in the least stored energy.

In the previous discussion of Fig. 3-8, the primary and secondary layers were in series. But the two primary layers and/or the two secondary layers could be paralleled, and the high frequency current would divide equally between the paralleled windings. The field must divide equally between the two winding portions in order to minimize the stored energy = ½ *BH*2volume. If the current and the field were to concentrate in one winding section, then in that one section *H* would double, and energy density would quadruple. Volume is halved, but net energy would double. Therefore current and field will balance nicely in both portions, for minimum energy and (coincidentally) minimum I^2R losses.

To achieve acceptable eddy current losses, it is often necessary to subdivide a wire whose diameter is greater than the penetration depth into many paralleled fine wires. Simply bundling these paralleled fine wires together won't do. Twisting the bundle won't help much. *Paralleled conductors within one winding section must all rotate through all levels of the winding,* so that each conductor has the same induced voltage integrated along its length. A special technique to ensure the proper division of current among the paralleled wires is used in the manufacture of **Litz Wire**, discussed in reference (R2).

Bear in mind that when a wire is subdivided into many fine wires to make the layer thickness smaller than D_{PEN}, the number of layers is correspondingly increased. For example, a single layer of solid wire replaced by a 10x10 array of 100 parallel fine wires becomes 10 layers when entering Fig. 3-5.

Passive losses

High ac losses can occur in windings that are carrying little or no current, if they are located in the region of high ac magnetic field intensity between primary and secondary. Situations of this nature include: Faraday shields, lightly loaded or unloaded secondaries, and the half of a center-tapped winding that is not conducting at the moment.

If the "passive winding" conductor thickness is not substantially less than D_{PEN}, the magnetic field cannot fully penetrate. Equal and opposite currents must then flow on opposite surfaces of the conductors in each layer of the passive winding to terminate or partially terminate the field on one side of each conductor and re-create it on the other side. Although the net current is zero, the surface currents can be quite high, causing significant additional winding loss.

Passive winding losses can be reduced or eliminated by:

- Relocating the winding out of the region of high ac field intensity.

- Reducing field intensity by interleaving and by using a core with wide window breadth.
- Making conductor thickness substantially less than D_{PEN}.

Faraday shields prevent electrostatic coupling between primary and secondary (more later). Of necessity they are situated where the field intensity is highest. Since they carry very little current, conductor thickness can and should be very much less than D_{PEN}.

With multiple secondaries, windings should be sequenced so the highest power secondary is closest to the primary. This keeps the lower-powered secondaries out of the highest field region, and has the added benefit of minimizing the adverse effect of leakage inductance on cross-regulation. This winding hierarchy is more difficult to achieve if the primary is interleaved outside of the secondaries. One way of accomplishing this, shown in Fig. 3-9, is to interleave the highest power secondary, S1, outside the lower power secondary(s), S2. The S1 sections (and/or the primary sections) can be either in series or parallel, whichever best suits the number of turns required.

Center-tap windings should be avoided. In center-tap windings, one side is inactive (passive) while the other side is conducting. Not only does this result in poor utilization of the available window area (compared with the single winding of the bridge configuration), the inactive side usually sits in the high field between the active side and the opposing windings, thereby incurring passive losses.

It is usually not difficult to avoid center-tap windings on the primary side, by choosing a forward converter, bridge, or half-bridge topology. But with low voltage secondaries, the importance of minimizing rectifier drops usually dictates the use of a center-tapped secondary winding.

Calculating rms Current

The relationships between peak current, I_P, total rms current, I, and its dc and ac components, I_{DC} and I_{AC}, are given below for several of the current wave-

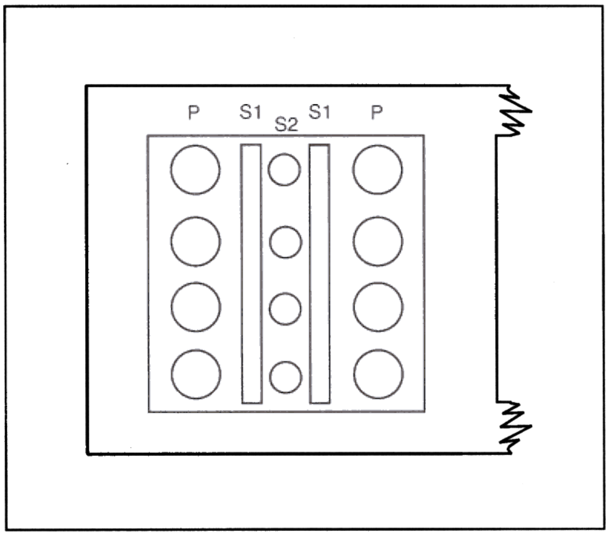
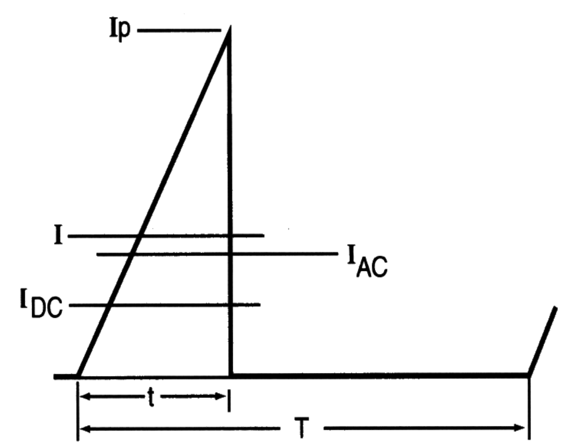


Figure 3-9 Interleaved Winding Hierarchy

shapes encountered in switching power supplies. For each half of a Center-Tap winding, which conducts only on alternate switching periods, substitute D/2 in place of D.

$$I^2 = I_{DC}^2 + I_{AC}^2 \qquad D = t/T$$

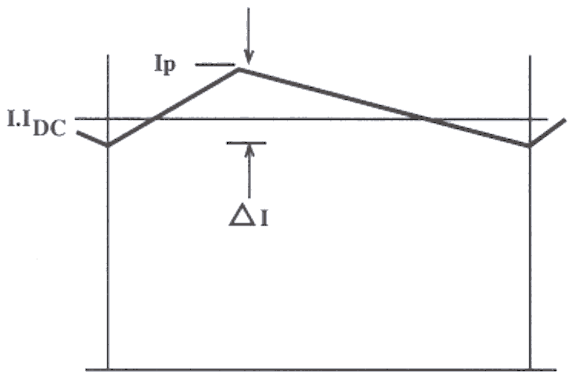


$$I_{DC} = I_P D/2$$

$$I_{AC} = I_P(D/3 - D^2/4)^{1/2}$$

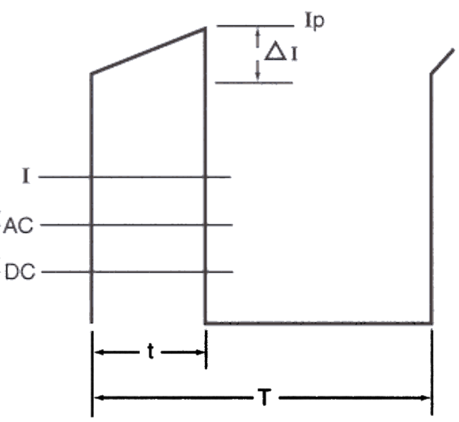
$$I = I_P(D/3)^{1/2}$$

Fig 3-10a Discontinuous Mode Waveform



$$I \approx I_{DC} = I_P - \Delta I/2$$
$$I_{AC} = \Delta I/(12)^{1/2}$$

Figure. 3-10b Continuous Mode -- Filter Inductor



$$I_{DC} = (I_P - \Delta I/2)D$$
$$I_{AC} = (I_P - \Delta I/2)[D(1-D)]^{1/2}$$
$$I = (I_P - \Delta I/2)D^{1/2}$$

Figure. 3-10c Continuous Mode -- Transformer

Window Utilization

How large a window is necessary to contain the ampere-turns of all the windings?. Ultimately, this is determined by maximum allowable power dissipation and/or temperature rise. A traditional rule-of-thumb for larger 60Hz transformers is to operate copper windings at a current density of 450 A/cm^2 (2900 A/in^2). However, smaller high frequency transformers can operate at much higher current densities because there is much more heat dissipating surface area in proportion to the heat generating volume.

How much of the window area is actually useful copper area? In a small transformer with a bobbin and with high voltage isolation requirements, perhaps only 25% or 30% of the window area is copper.

Isolation safety requirements dictated by specs such as IEC 65 and VDE 0860 impose 3 layers of insulation between windings with minimum creepage distances of 6 to 8mm from primary to secondaries around the ends of the insulation layers at both ends of the winding. Nearly 1 cm of window breadth is lost to this requirement, severely impacting window utilization, especially with small cores in low power transformers. The increased separation also results in higher leakage inductance between primary and secondary.

Recent revisions to these specifications permit the use of triple-insulated wire (more about this later), which can reduce the penalty imposed by isolation requirements. This is an area where expert advice can be very worthwhile.

A bobbin, if used, significantly reduces the available window area, again impacting smaller, low power applications much more heavily.

Voids between round wires waste 21% of the winding cross-section area. Wire insulation further reduces the useful area, especially with the smaller diameter wires used to minimize high frequency losses, because insulation is a larger percentage of small wire diameter.

In Litz wire, with multiple levels of twisting very fine insulated wires, the amount of copper may be less than 30% of the total cross-section. There is no benefit in pushing to an R_{AC}/R_{DC} of 1.5, if R_{DC} is increased two or three times because of the reduced copper area of the many fine wires involved.

Topology considerations: Bridge and half-bridge buck-derived topologies have the best primary winding utilization because the entire winding conducts most of the time. Forward converter winding utilization is less efficient, because the windings usually conduct less than 50% of the time. Higher peak primary current is required for the same power output. Center-tapped windings utilize window area inefficiently because half of the winding is inactive while the other half conducts.

To balance the losses between primary and secondary(s), primary and secondary copper areas should be approximately equal for forward converters and for Center-tapped primaries with center-tapped secondaries. For bridge or half-bridge primary with center-tapped secondaries, the primary copper should be approximately 40%, with the secondaries totaling 60%.

Triple-Insulated Wire

Agency safety requirements for input-output isolation have imposed a severe burden on high frequency transformer design. The purpose of high frequency operation is to reduce size, weight and cost, but creepage and clearance requirements essentially waste almost 1 cm of the winding breadth This can make the transformer considerably larger than it would otherwise be, especially with small transformers in low power applications.

Recently, the ability to extrude three layers of insulation over a single wire or over an entire Litz wire bundle holds out the possibility of reducing the substantial penalty incurred by creepage and clearance requirements, thus enabling significant size reduction. Individual conductors in a triple-insulated Litz cable are single insulated, minimizing the penalty of the thicker triple layers.

Although triple-insulated wire is much more costly than conventional magnet wire or Litz wire, reduced transformer size would make it cost-effective.

Teflon FEP is presently approved as an insulation material, and other materials are under review. As of 1997, agency specifications are undergoing revision, and the situation is quite volatile. The author has received conflicting opinions regarding the usability of triple-insulated wire. The transformer designer should consult an engineer conversant with the situation for guidance *early in the design phase.*

Some of the vendors who indicate they can supply triple-insulation wire include:

Rubadue Wire Co., Inc. (CA)	714-693-5512
Furukawa Electric (GA)	770-487-1234
New England Elec. Wire (NH)	603-838-6625
Kerrigan Lewis Wire Prod. (IL)	773-772-7208

Optimized Winding Placement

A fundamental principle of magnetic device design is not to allow the total magnetic force, ι, build up to a substantial level. In a transformer, the magnetic source is the ampere-turns of the primary winding. The magnetic "load" is the nearly equal and opposite ampere-turns of the secondary(s). If the primary winding is put on one leg of a simple C-core, and the secondaries across the other leg, then the full magnetic force appears across the two core halves, radiating considerable stray flux to the outside world, resulting in high EMI and high leakage inductance. But if the secondary winding conforms to the primary, i.e., is wound directly over the primary on the same core leg, then the ampere-turns introduced by the primary are offset by the secondary ampere-turns, turn for turn, and the total magnetic force never builds to a substantial value. There is almost zero magnetic potential across the core halves which act simply as a short-circuit return path for the flux. There is very little stray flux, and leakage inductance is small. On a toroidal core, all windings should be uniformly distributed around the entire core.

With an inductor or flyback transformer, the magnetic source is the primary winding, the "load" is the gap where the energy is stored. With a gapped ferrite core, put the winding directly over the gap. Better, since the winding is distributed along the center-leg, distribute the gap along the center-leg with two or three correspondingly smaller gaps. If the gap is distributed, as in a powdered metal toriod, then the winding should be uniformly distributed around the core.

Ideally, the magnetic source and load should be as intimate and conformal as possible, being distributed in exactly the same way.

A ferrite toroid with a discrete gap does not conform to a winding distributed around the toroid. This is a complete disaster in terms of stray flux.

Inter-Winding Capacitance

Inter-winding capacitance causes feedthrough of common mode noise from the switched primary winding to the secondary. Unfortunately, techniques that reduce leakage inductance and eddy current losses–interleaving, broad window, close P to S spacing–cause increased interwinding capacitance.

The proper use of Faraday shields will reduce the adverse coupling of the interwinding capacitance. A Faraday shield is an electrostatic shield consisting of thin foil or metalized insulating film wrapped completely around the interwinding space between primary and secondary. Where the shield overlaps, it must be insulated from itself to avoid being a shorted turn. The shield should be much thinner than the penetration depth D_{PEN} to avoid passive eddy current losses.

The shield should be tied directly to the quiet side of the transformer primary, with minimum lead inductance. It should *not* be grounded, or much common-mode EMI will be propagated to the input.[5] Primary to shield capacitive charging current will add to losses in the primary-side switch unless ZVT techniques are used.

End-to-end capacitance

End-to-end capacitance, sometimes called "distributed capacitance", appears in shunt with a winding. In a transformer, this capacitance results in series and parallel resonances with mutual inductance and leakage inductances. In a filter inductor, the end-to-end capacitance, above resonance, passes the high frequency components of the switched waveform through to the output.

End-to-end capacitance has a negligible effect in low voltage, low impedance windings, but in high voltage windings, it is a serious problem. High voltage windings have many turns which are usually wound back and forth in multiple layers. Thus the end of the second layer is directly over the beginning of the first layer. The effect of the significant capacitance between these layers is magnified because of the large ac voltage across the many intervening turns.

A single layer winding will have very little end-to-end capacitance, although there is a possible sneak path from end to core to end, unless the core is tied to an ac quiet point.

In a multi-layer winding, end-to-end capacitance is reduced dramatically by sectionalizing the winding along the available length–a technique often used in RF chokes, probably impractical in SMPS magnetics.

The "bank winding" technique for reducing end-to-end capacitance is shown in Figure 3-11.

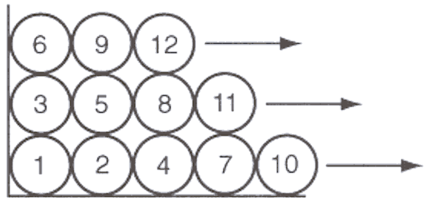


Figure. 3-11 Bank Winding

With bank winding, the capacitance between physically adjacent turns has little effect because there are few electrically intervening turns, thus low voltage across the capacitance, compared with the conventional "back and forth" winding technique.

Points to Remember

- Paralleling windings or wires within windings succeeds only if the expected division of high frequency current results in the smallest energy transfer.
- Skin Effect: A single wire has an infinite number of parallel paths within itself. High frequency current flows in the paths near the surface and not the paths in the center because this minimizes energy expenditure.
- Proximity effect: In a pair of conductors or windings thicker than D_{PEN}, with opposing currents, high frequency current flows only on those surfaces closest to each other, and will spread across those surfaces, in order to minimize expended energy.
- If multiple layers are connected in parallel, HF current will flow only on the inner surface of the inner layer. If the layers are connected in series, the same current must flow in all layers, but if layer thickness is greater than D_{PEN}, opposing high frequency currents of great magnitude will flow on the inner and outer surfaces of each layer,

Section 4 Power Transformer Design

Power Transformer Design

This Section covers the design of power transformers used in buck-derived topologies: forward converter, bridge, half-bridge, and full-wave center-tap. Flyback transformers (actually coupled inductors) are covered in a later Section. For more specialized applications, the principles discussed herein will generally apply.

Functions of a Transformer

The purpose of a power transformer in Switch-Mode Power Supplies is to transfer power efficiently and instantaneously from an external electrical source to an external load. In doing so, the transformer also provides important additional capabilities:

- The primary to secondary turns ratio can be established to efficiently accommodate widely different input/output voltage levels.
- Multiple secondaries with different numbers of turns can be used to achieve multiple outputs at different voltage levels.
- Separate primary and secondary windings facilitate high voltage input/output isolation, especially important for safety in off-line applications.

Energy Storage in a Transformer

Ideally, a transformer stores no energy–all energy is transferred instantaneously from input to output. In practice, all transformers do store some undesired energy:

- Leakage inductance represents energy stored in the non-magnetic regions between windings, caused by imperfect flux coupling. In the equivalent electrical circuit, leakage inductance is in series with the windings, and the stored energy is proportional to load current squared.
- Mutual inductance (magnetizing inductance) represents energy stored in the finite permeability of the magnetic core and in small gaps where the core halves come together. In the equivalent circuit, mutual inductance appears in parallel with the windings. The energy stored is a function of the volt-seconds per turn applied to the windings and is independent of load current.

Undesirable Effects of Energy Storage

Leakage inductance delays the transfer of current between switches and rectifiers during switching transitions. These delays, proportional to load current, are the main cause of regulation and cross regulation problems. Reference (R4) included in this manual explains this in detail.

Mutual inductance and leakage inductance energy causes voltage spikes during switching transitions resulting in EMI and damage or destruction of switches and rectifiers. Protective snubbers and clamps are required. The stored energy then ends up as loss in the snubbers or clamps. If the loss is excessive, non-dissipative snubber circuits (more complex) must be used in order to reclaim most of this energy.

Leakage and mutual inductance energy is sometimes put to good use in zero voltage transition (ZVT) circuits. This requires caution–leakage inductance energy disappears at light load, and mutual inductance energy is often unpredictable, depending on factors like how well the core halves are mated together.

Losses and Temperature Rise

Transformer loss is sometimes limited directly by the need to achieve a required overall power supply efficiency. More often, transformer losses are limited by a maximum "hot spot" temperature rise at the core surface inside the center of the windings. Temperature rise (°C) equals thermal resistance (°C/Watt) times power loss (Watts).

$$\Delta T = R_T \times P_l$$

Ultimately, the appropriate core size for the application is the smallest core that will handle the required power with losses that are acceptable in terms of transformer temperature rise or power supply efficiency.