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.

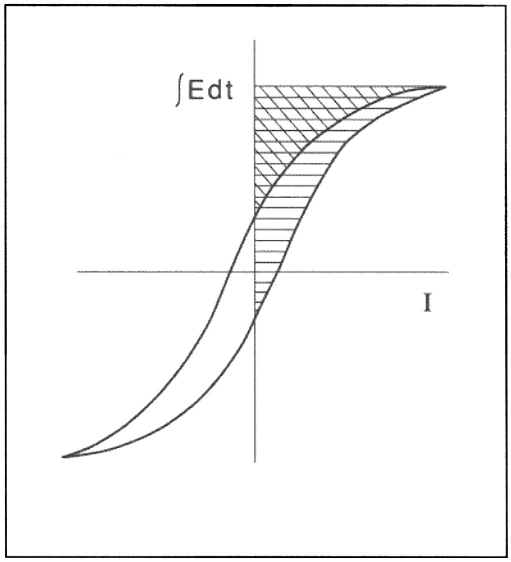

Figure. 2-1 Ferrite Core Characteristic

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.

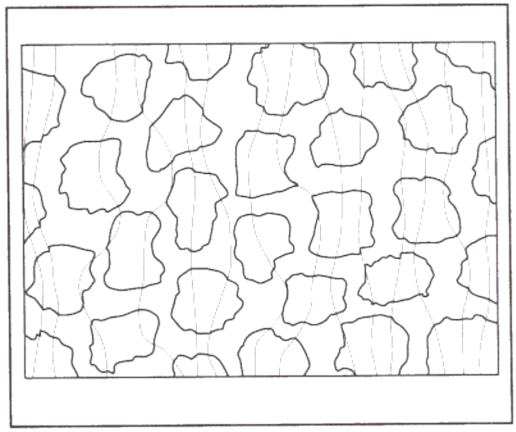

Figure. 2-2 -- Easy Flux Path Between Particles

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 $\int E dt$ = applied Volt-seconds, N = turns, and A_e = core cross-section area:

$$\Delta B = \frac{1}{NA_e} \int E dt$$

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.

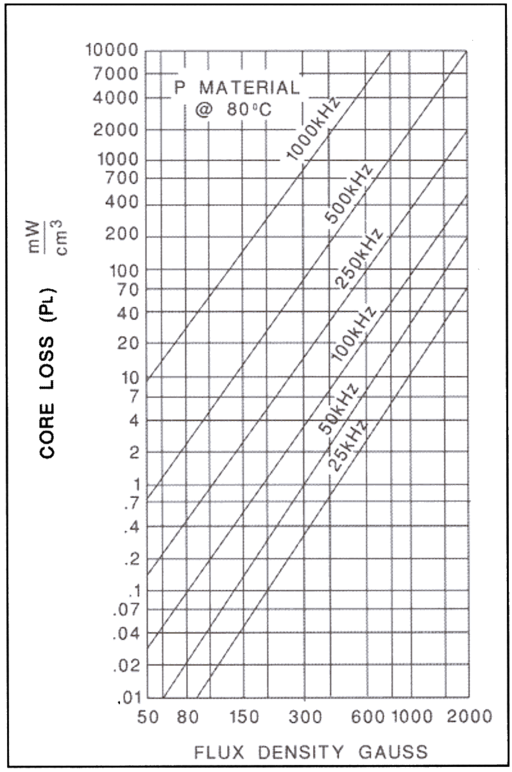

Figure. 2-3 Core Loss -- "P" Material

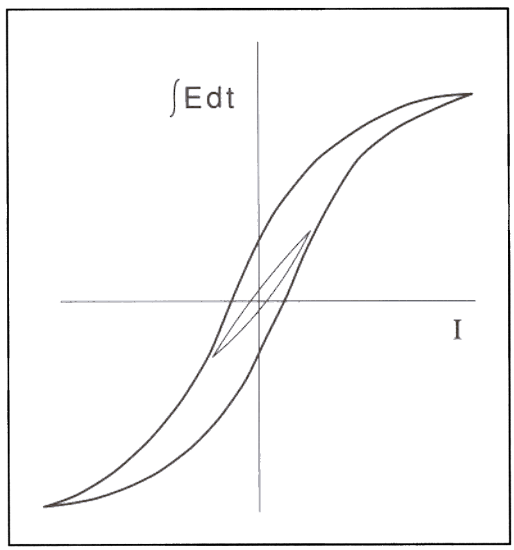

Figure. 2-4 – Minor Hysteresis Loop

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.

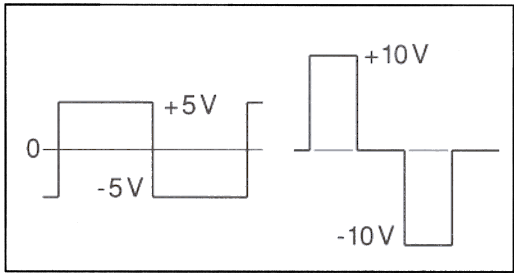

Figure. 2-5 Waveform Comparison

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.

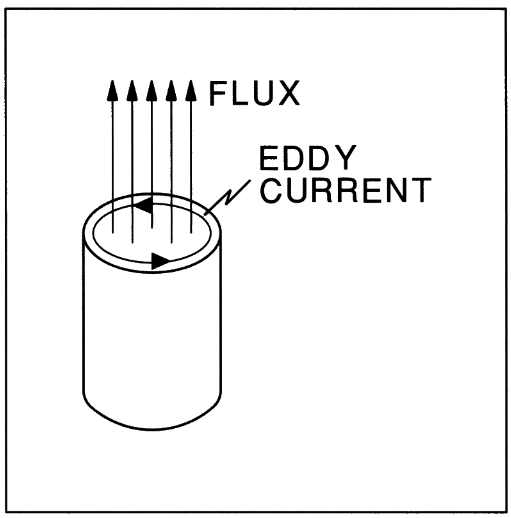

Fig. 2-6 Eddy Current in a Ferrite Core

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

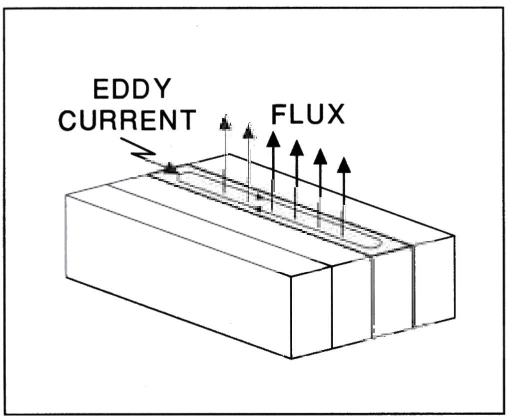

Figure. 2-7 Lamination Detail

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).

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, $\rho = 55 \bullet 10^{-4}$ 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 B_{SAT} 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.

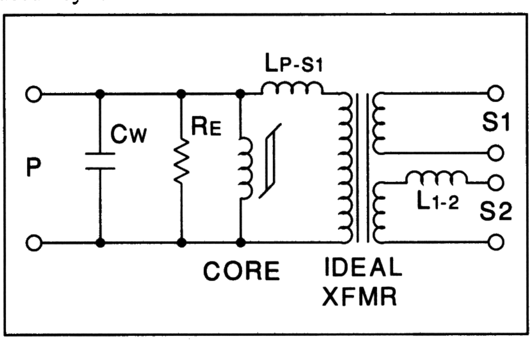

Fig. 2-8 Transformer Model

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

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