From Powder to Transformer

Majid Dadafshar

Abstract — Magnetic design is a critical element in designing any switch-mode power supply. Understanding the process of making magnetic materials from powder into a final core shape, and understanding the effects of parameters such as core loss, permeability, and flux density, designers are better able to select the right material to achieve the smallest size with lowest losses and reduced cost. By understanding AC losses and layer arrangements, designers can also optimize the windings to create a small and highly efficient magnetic design.

I. INTRODUCTION

Many designers view the pulse width modulator (PWM) as the brain and the magnetic element as the heart of the switching power supply. Understanding magnetic materials and their effect on the performance of SMPSs can be a powerful complement to the performance of any PWM controller. The intent of this paper is to convey a good understanding of magnetic material types, the process they go through, and their relation to power supply performance.

II. FUNDAMENTALS

A. *Magnetic Materials and Their Manufacturing Processes*

Magnetic materials consist of individual molecular magnets, capable of movement within material. These magnetic particles are arranged randomly when unmagnetized, with a net flux of zero outside the material. When they are placed in the vicinity of an external field, they are magnetized and each particle aligns with the field in a definite direction. The degree of magnetization depends on the degree of the alignment of particles; when all the particles are aligned, the material is saturated.

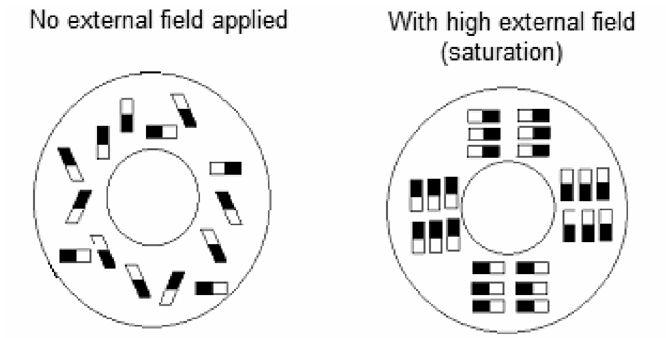

Figure 1. Individual magnets' alignment with and without external fields.

Ferrites are commonly used for switched-mode power supply designs and the magnetic material composition is an oxide ceramic usually containing Fe and additional materials like Mn, Ni and Zn.

There are three major processing steps in making a ferrite:

1. Composition, batching, and mixing
2. Milling, spray drying, pressing, and sintering
3. Grinding, coating, and testing.

During the composition process, the raw material impurities must be kept low, with a high purity level of 99.5% for Fe_2O_3. Batching helps correct weighting and mixing creates a uniform distribution.

Milling yields powder with controlled particle size for low power loss and spray drying helps create a spherical powder. The pressing density must be controlled to achieve desired sintering dimensions, as sintering shrinkage is a function of the density of the unsintered part.

During the grinding process, proper control of the flatness and mating surfaces is achieved, which corresponds to $A_L=L/N^2$ values for each core set. After these processes, a new core set can be created.

B. *Ferrites*

Ferrites are commonly used in power supply designs and are categorized into either hard or soft type materials. Hard materials, like permanent magnets, are very difficult to magnetize without a high degree of coercive force (H_c). Soft materials, like ferrites, can easily be magnetized with small coercive force.

The soft ferrites subdivide into two commonly used in switched-mode power supply designs.

1. *MnZn (Manganese Zinc)*
   - Composed of 50% Iron oxide (Fe_2O_3) + 30% Mn_3O_4+20% ZnO
   - High Permeability (μ_i) of 2300-2800 Range
   - Max. Flux Density (B_m) of 4200G-4900G
   - Tailored for Frequency Range of <3MHz
   - Low Resistivity (50-2000Ω-cm)

2. *NiZn (Nickel Zinc)*
   - Composed of 50% iron oxide (Fe_2O_3) + 30% NiO+20% ZnO
   - Low Permeability (μ_i) of 250-375 Range
   - Max. Flux Density (B_m) of 2000G-4500G
   - Tailored for High Frequency >3MHz
   - High Resistivity (10E5 – 10E9)

In higher-frequency applications using *NiZn* material, the higher resistivity translates into lower eddy current losses with a penalty in higher magnetizing current due to lower permeability or inductance.

The Curie temperature, saturation flux density, and permeability can be changed to different values by changing the percentage and the ratio of the iron oxide (Fe_2O_3, 50%-56%) versus MnO (27-38%). Based on the frequency of operation and efficiency, different materials are needed for specific applications.

The Curie temperature is the point where the ferromagnetic materials lose their magnetic characteristics.

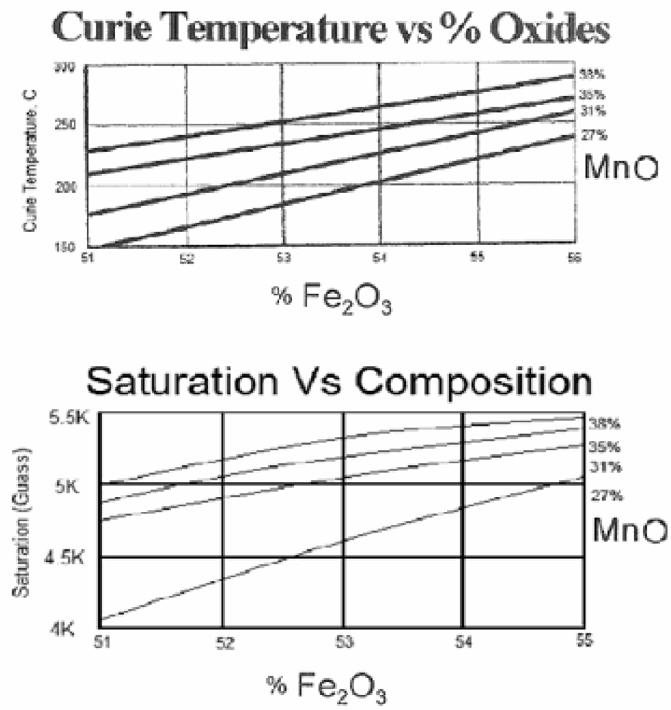

Figure 2. Effect of changing % of oxides and composition in material characteristic.

These changes allow the core suppliers to provide a wide variety of materials with different saturation flux density or core loss levels. For example, in continuous mode operation, with small ripple current and flux swing, having a material with higher core loss but greater saturation flux density (powder iron) is desirable. On the other hand, in discontinuous mode applications, where there is a large flux swing, a material with lower core loss (ferrite) is desirable.

C. Magnetism

Electricity and magnetism are very closely related and Maxwell equations (Faraday, Ampere, and Lenz laws) can be used to create formulas to determine magnetic relations and to design magnetic elements.

Faraday's law states that a magnetic field is a result of an electrical current and the flux through any winding should equal the integral V-µs per turn. Once a conductor is placed in a time-varying magnetic field, electromotive force (EMF), proportional to the rate of change of magnetic flux, is induced in the conductor, causing a current flow. This can be stated as:

$$V = -N\frac{d\phi}{dt} \tag{1}$$

where:
V = Induced EMF (V)
N = Number of turns on winding
t = Time (seconds)
ϕ = Magnetic flux (unit: Wb)

Since the magnetic flux is contained within a given core size of cross-sectional area (A), the magnetic flux density (B), can be defined as:

$$B = \Phi / A \quad (\mathrm{Wb/m^2}) \tag{2}$$

Amperes law states that the applied magnetomotive force or ampere-turns (NI) is equal to the sum of products of discrete magnetizing force in A/m (H) and flux path length in meter (1e) around any closed magnetic circuit. This is shown by:

$$\sum NI = \oint H \bullet dl_e = H_m l_m + H_g . l_g \tag{3}$$

In the same way that voltage causes a current flow, the magnetizing force (H) generates lines of flux. The flux density depends on the magnitude of the magnetizing force and the core material.

From Faraday's law:

$$V = N\frac{d\phi}{dt} = NA\frac{dB}{dt} = NA\mu\frac{dH}{dt} \tag{4}$$

From Ampere's law:

$$NI = Hl_e \tag{5}$$

Combining the two gives:

$$V = \frac{N^2 A\mu}{l_e}\frac{di}{dt} \tag{6}$$

or:

$$V = \left[\frac{N^2 A\mu}{l_e}\right]\frac{di}{dt} = L\frac{di}{dt} \tag{7}$$

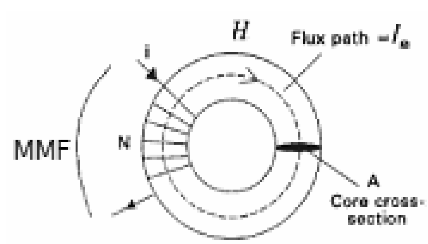

Figure 3. Example of a magnetic circuit.

D. BH Loop

When power materials are made into different shapes with a given effective cross-sectional area (A_e) and magnetic path length (L_e), the BH loop characterizing the material can be translated into magnetic flux $\Phi = B\ A_e$ versus magnetomotive force $F = H\ L_e$, as shown in Figure 4. When the turns are added to the core shapes, electrical devices (inductors or transformers) are created that are characterized by voltages, currents, and inductances.

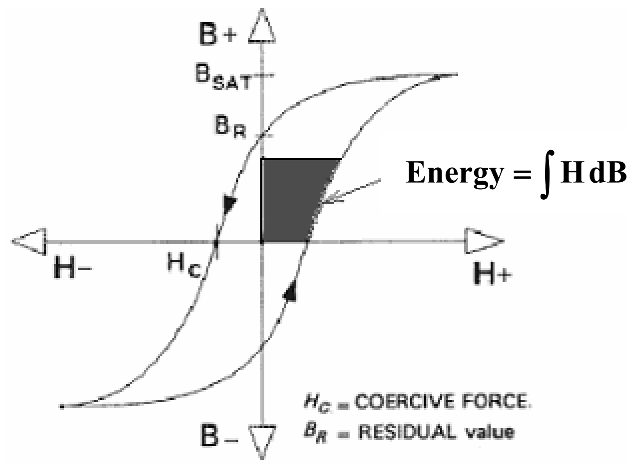

Figure 4. Material BH loop.

The slope of the BH loop indicates initial permeability. Notice the slope in Figure 4 is constant for only small increments in B and H, where H is the forcing function and B is the result. As H increases, a point is reached at which B no longer increases and the core is saturated (B_{sat}). At this saturation point, the material becomes non-magnetic and is no longer a low-reluctance flux path.

The BH loop characterizing the material can be translated into ($B \bullet A_e$) and ($H \bullet l_e$), when powder materials are made into different shapes, with a given cross-sectional area (A_e) and magnetic path length (l_e). When turns are added into the core shape, an electrical equivalent of the core is available in voltage, current, and inductance.

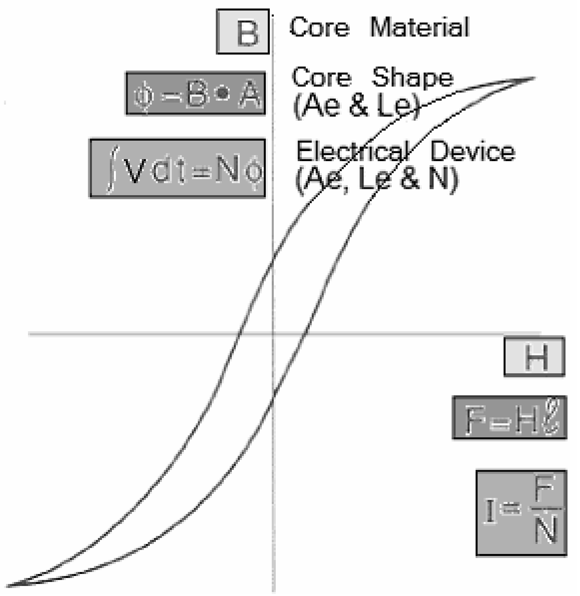

Figure 5. Transformation of material to electrical equivalent.

III. CORES

A. Core Shapes

Cores provide a low-reluctance flux path and come in different shapes and sizes. Each shape has a different window configuration, some with a wide window area to minimize the number of layers, leakage, and AC resistance (R_{AC}).

Pot cores, for example, have good magnetic shielding but very small window area, not suited for AC-DC applications where the creepage allowance would waste a large portion of the window area. Also, since all the copper is surrounded by the core area, it should be designed to have lower copper loss compared to core loss.

EE core, on the other hand, has a larger window area relative to core size and can have an equal ratio between copper and core loss. Toroids have their core covered by the windings and need to have a higher copper loss compared to core loss.

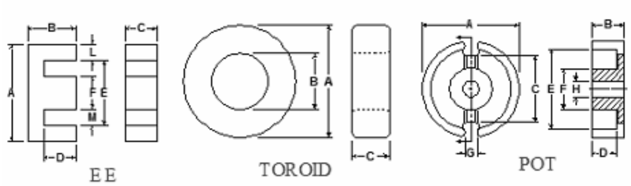

Figure 6. Different core shapes.

Once the migration from powder-grain size to an actual-core size has occurred, the material's cross-sectional area and magnetic path length are associated. Understanding the B-H curve is a basic tool for understanding magnetic characteristics.

K is the window utilization factor and in the majority of AC to DC applications, safety isolation requirements for creepage and clearance can easily consume more than 36% of the available window area, causing a reduction in window utilization. Finding a core shape with a wide window and lower height lowers the impact of fixed creepage, causing better widow area utilization.

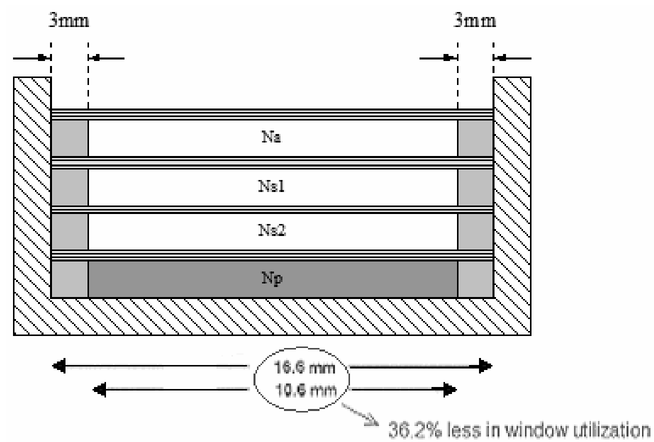

Figure 7. Window utilization for safety requirement.

B. Core Power Handling Capability

Since a major cost of any SMPS is the transformer, optimal size can reduce the size and cost of the power supply. For a chosen operating frequency, the total power handling capability of any core (P_{total}) is a function of flux density (B_{pk}), current density of conductors (C), core cross-sectional area (A_e), bobbin winding area (A_b), and whether the core is being magnetized bi-directional or uni-directional.

The choice in using a uni-directional (flyback) or bi-directional (push-pull) magnetization or topology has a significant impact on the transformer design and size for any given output power.

The following equation determines an initial and relatively accurate core size:

$$A_e A_b = \frac{P_{total}\, 10^8\, Fk}{4\, \eta\, B_{pk}\, f\, K\, C\, \phi UF} \qquad (8)$$

where:

η = Efficiency
B_{pk} = Peak AC flux density (Gauss)
f = Frequency (Hz)
K = Window utilization Factor
A_e = Core cross-sectional area (cm^2)
A_b = Bobbin window area (cm^2)
C = Current density (circular mil/Amp)
ϕUF = Magnetization (1=unidirection, 2=bidirection)
Fk = Cooling factor

By changing any of the parameters in Equation 8, a smaller core can be used for any given output power. Figure 8 shows the effect of three different flux densities in the total power handling capability (P_{total}) for a given core size.

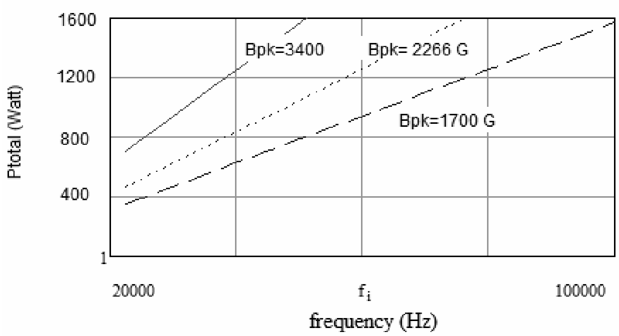

Figure 8. Effect of flux density in core power handling.

C. Gap

The amount of energy stored within the magnetic element can be increased by installing a gap to reduce the slope of the B-H loop and effective permeability, causing an increase in the energy storage by allowing additional energy to be stored in the air gap. Although the gap is undesirable in transformer applications (except ZVS phase shifted FB), it is desirable in inductor or flyback applications. The gap size can be found for a given core cross-sectional area (A_e) and needed inductance (L) through:

$$l_g = \frac{\left(4\pi \times 10^{-7}\ \mathrm{H/m}\right) N^2 A_e}{L} \tag{9}$$

D. Flux Density & Core Loss

The combined copper and core losses are often times determined by the required overall power supply efficiency and temperature rise of the transformer.

The core loss (P_c) represents the energy lost within the ferromagnetic material as the dipoles are rotated back and forth with the alternating magnetic field. This loss is a function of the amount of rotation (flux swing), the frequency of rotation (frequency), and the volume of the core material (V_c).

$$P_c = C_m B_{ac}^{y} Freq^{x} V_c \quad (\mathrm{mW/cm^3}) \tag{10}$$

C_m, y, and x are material constants provided by core suppliers. Table I below shows curve-fitted power loss results for common materials.

Flux density should be chosen for acceptable core loss and to keep the core below the saturation.

TABLE I
CORE CONSTANTS FOR DIFFERENT FERROXCUBE MATERIALS AT DIFFERENT FREQUENCIES

Grade	Frequency (kHz)	cm	X	Y
3C80	10-100	16.7	1.3	2.5
3C81	10-100	7	1.4	2.5
3C85	20-100	11	1.3	2.5
	100-200	1.5	1.5	2.6
3F3	20-300	.25	1.6	2.5
	300-500	2.10^{-2}	1.8	2.5
	500-1000	36.10^{-7}	2.40	2.25
3F4	500-1000	12.10^{-2}	1.75	2.90
	1000-3000	11.10^{-9}	2.8	2.4

Core geometry and the primary number of turns are chosen to keep the flux density low enough to prevent the core from saturating, but high enough to efficiently use the magnetic capabilities of the core material. The power losses, like permeability, vary with temperature. Using a core-loss density of around 220mW/cm^3 for a 40-degree rise should be a safe design assumption.

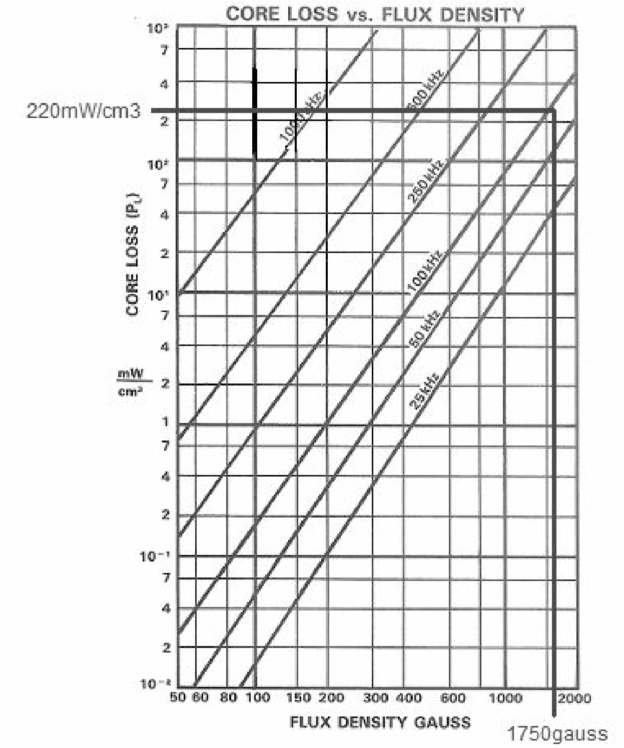

Figure 9. Core loss for any flux density at an operating frequency.

If a higher frequency is desirable, lower the flux density to maintain 220mW/cm^3 or keep the same flux density and increase the product of the core and bobbin window area (A_eA_b).

IV. WINDINGS

A. DC and AC Losses

Low frequency and DC resistance (R_{DC}) losses can be calculated from winding dimensions, but high frequency and AC resistance (R_{AC}) losses are more difficult to calculate accurately. To have a more accurate power loss determination, the AC resistance can be found by multiplying R_{DC} by a correction factor ($Fr=R_{AC}/R_{DC}$), and calculated using Dowell's equation [5]:

$$Fr(m,x) := (skin) + (proximity) \tag{11}$$

$$Fr(m,x) := \frac{x}{2} \bullet \frac{\sinh(x)+\sin(x)}{\cosh(x)-\cos(x)} + (2 \bullet m - 1)^2 \bullet \frac{x}{2} \bullet \frac{\sinh(x)+\sin(x)}{\cosh(x)-\cos(x)}$$

where:

x = conductor height/skin depth, and

m = number of layers in winding section from MMF=0 to MMF= maximum.

The ultimate goal in selecting proper core size, wire size, and current density is the temperature rise of the transformer. To predict a more accurate loss, it is necessary to understand the AC losses (skin, proximity).

B. Skin, Proximity, and Fringing Losses

The AC losses can be divided into three main categories: skin, proximity, and fringing.

1. Skin effect is caused by eddy currents within a single conductor. It is not changed by layer arrangement.

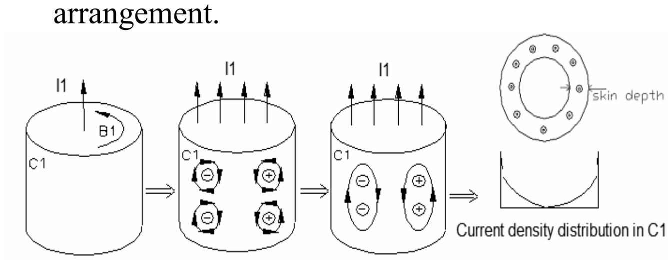

Figure 10. Skin effect.

When the current I1 flows through the conductor C1, the eddy current caused by internal field B1 forces the current distribution to concentrate on the surface and reduces the effective copper area, increasing the AC resistance.

The skin depth in meters can be calculated for any frequency (Hz) with the equation:

$$\delta = \sqrt{\frac{2}{\omega\,\mu\,\sigma}} = \frac{0.065\,\text{m}}{(f\ \text{in Hz})} \tag{12}$$

$\omega = 2 \bullet \pi \bullet f$ Angular frequency of current

$\mu = \mu_o \bullet \mu_r$ Absolute permeability of conductor

μ_o = Permeability of air, $4\pi \bullet 10^{-7} (H/m)$

μ_r = Relative permeability of conductor

σ Electrical conductivity, Copper = $59.6 \cdot 10^6$ (S/m)

2. Proximity effect is caused by eddy currents induced in a conductor by the current in an adjacent conductor. It is changed by layer arrangement.

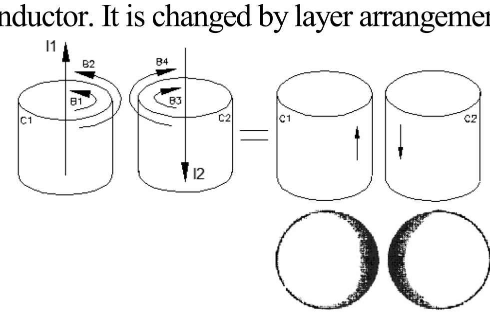

Figure 11. Proximity effect.

3. When magnetizing inductance is set by inserting an air gap in flyback transformers, flux is no longer confined to the core. Flux begins to fringe as the gap is increased, so that the effective gap area is larger than the cross-sectional area of the core.

   The gap-fringing field can cause significant eddy current loss, which needs to be added to the main copper loss calculation.

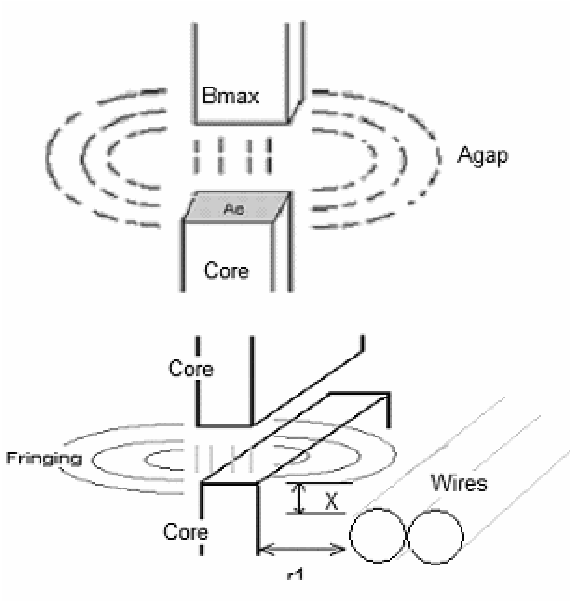

Figure 12. Fringing field.

By proper separation between distances r1 and X, the fringing field effect can be minimized. Keep in mind that increasing r1 reduces the much-needed window area. By changing X, this changes the normal field to a tangential field and reduces the effect of the net flux.

Any winding turns in the vicinity of the high flux density of the fringing field (flyback gap) can experience large eddy current losses. Try to design the windings such that there is no copper in the fringing field.

Proximity losses can be significantly higher than the skin effect losses. Proper layer arrangement, using an MMF diagram, can minimize the proximity effect.

C. Magnetomotive Force (MMF) Diagrams & Layer Arrangement

MMF diagrams are a powerful tool for designing correct layer arrangements to minimize proximity loss (AC copper losses) and lower leakage inductance saving snubber loss and lowering emissions.

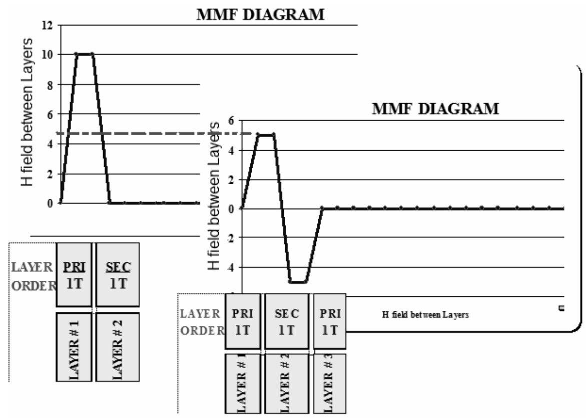

Figure 13. MMF diagram for layer arrangement.

As windings are built into layers, the proximity to each other further crowds the current and increases the resistance. The harmonic AC resistance is higher as the number of layers increases. Since current crowding is caused by the magnetic field intensity, interleaving can be helpful. Interleaving reduces the effective number of layers in a winding portion, thereby reducing the H field between the layers.

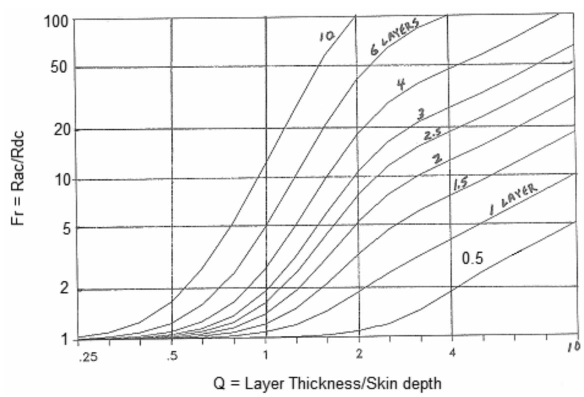

Figure 14. Dowell's equation in calculating AC resistance for different number of windings layer.

Leakage inductance is the imperfect magnetic coupling between the primary and secondary windings and can be stored between layers. Interleaving causes a reduction in leakage inductance and field intensity (H) buildup between each layer with a penalty of higher intra-winding capacitive coupling of common-mode noise between primary and secondary windings.

D. Current Density

Current density has a direct effect in winding copper losses and load regulation. It should be chosen to minimize the output voltage variation for a load change or to lower copper losses.

Using current density C values in the 250-550 circular mils per amp range (J=4-8 Amp/mm^2) should be a safe number to use, depending on the thermal design of the transformer. For a given peak primary or secondary current and a chosen current density, the area in circular mils is $I_p \bullet C$ (in mm^2, it is I_p/J). Using a magnetic wire table, the closest wire size (AWG) can be selected.

V. TEMPERATURE RISE

A. Temperature Rise

Once the core losses, copper losses and AC losses are calculated, estimate the temperature rise (degrees C) of the transformer using the following empirical formula:

$$\Delta T := Fk\left[50.6 \bullet \left(\frac{P_{core}}{P_{cu}}\right) + 88.1\right] \bullet \frac{P_{cu}}{As^{.853}} \tag{13}$$

where
As = Surface area of the core/coil (in^2)
Fk = Forced air cooling factor
Power in watts

Since the transformer has many thermal resistance paths between core and the windings to the transformer surface (with many of them unknown), accurate calculation of the temperature rise is extremely difficult. Equation 13 is only an estimate.

Both copper and core losses must be within a certain limit to achieve power supply efficiency and to generate acceptable temperature rise for a given core size.

Figure 15 shows the change in the cooling factor (Fk), in the vertical axis, against the rate of air flow in the horizontal axis. Increasing the operating peak flux density, current density in the winding, or applying more air flow can reduce the size of the selected core with an increase in losses.

Consider using 1.0 for no airflow, 0.7 for 40 LFM (linear feet per minute), 0.45 for 120 LFM, and 0.35 for 240 LFM.

CFM relates to volume and LFM relates to velocity and, since most engineers deal with LFM, use Equation 14 to change the CFM to LFM:

$$Velocity(LFM) = Volume(CFM) / Area(ft^2) \tag{14}$$

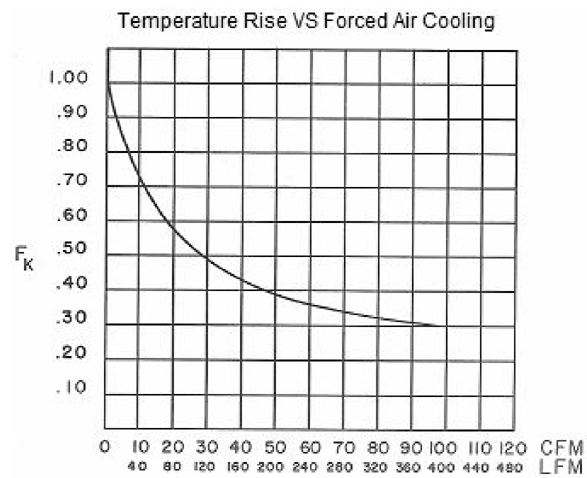

Figure 15. Cooling factor for different LFM and CFM assuming 0.25 ft^2 cross section for air flow.

VI. CONCLUSION

Designing transformers for switched-mode power supplies is a process of selecting the best core material, core shape, correct winding layer arrangements, and many other key specifications. All key parameters, like regulation, efficiency, size, cost, and electromagnetic interference (EMI) behavior of the supply are highly dependent on how the transformer is designed. The transformer is like a black box that needs to be opened and each part of this box needs to be well-understood to create an effective power supply design.

Understanding ferrite materials and their variations helps designers choose the correct material with the lowest core losses.

Proper selection of the window shape of the core can minimize the number of layers and effectively lower the AC winding losses and leakage.

Examining the total DC copper losses and AC losses (skin and proximity) helps the designer determine a more accurate loss estimate to meet the overall power-supply efficiency and provides a more accurate temperature-rise prediction.

Working with a magnetic supplier who is familiar with all issues related to core material and losses associated within a core and with a strong understanding of both DC and AC losses is important in designing a transformer.

VII. References

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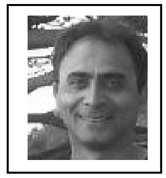

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