Temperature Rise Limit

In consumer or industrial applications, a transformer temperature rise of 40-50°C may be acceptable, resulting in a maximum internal temperature of 100°C. However, it may be wiser to use the next size larger core to obtain reduced temperature rise and reduced losses for better power supply efficiency.

Losses

Losses are difficult to predict with accuracy. Core loss data from core manufacturers is not always dependable, partly because measurements are made under sinusoidal drive conditions. Low frequency winding losses are easy to calculate, but high frequency eddy current losses are difficult to determine accurately, because of the high frequency harmonic content of the switched rectangular current waveshape. Section 3 discusses this problem extensively. Computer software can greatly ease the difficulty of calculating the winding losses, including high order harmonics(1).

Thermal Resistance

Temperature rise depends not only upon transformer losses, but also upon the thermal resistance, R_T (°C/Watt), from the external ambient to the central hot spot. Thermal resistance is a key parameter, unfortunately very difficult to define with a reasonable degree of accuracy. It has two main components: internal thermal resistance R_I between the heat sources (core and windings) and the transformer surface, and the external thermal resistance R_E from the surface to the external ambient.

Internal thermal resistance depends greatly upon the physical construction. It is difficult to quantify because the heat sources are distributed throughout the transformer. R_I from surface to internal hot spot is not relevant because very little heat is actually generated at that point. Most of the heat generated in the core (other than in toroids) is near the transformer surface. Heat generated within the winding is distributed from the surface to the internal core. Although copper has very low thermal resistance, electrical insulation and voids raises the R_T within the winding. This is a design area where expertise and experience is very helpful. Fortunately, internal thermal resistance is considerably smaller than external R_E (except with high velocity forced air cooling), and while R_I shouldn't be ignored, it usually is not critically important compared with R_E.

External R_E is mainly a function of air convection across the surface of the transformer–either natural convection or forced air. R_E with natural convection cooling depends greatly upon how the transformer is mounted and impediments to air flow in its vicinity. A transformer mounted on a horizontal surface and surrounded by tall components, or mounted in a relatively small enclosure will have considerably greater R_E than if it were mounted on a vertical surface, benefiting from the "chimney effect". With forced air cooling, R_E can be driven down to a very small value, depending on air velocity, in which case internal R_I becomes the primary concern. With forced air cooling, thermal resistance and temperature rise often become irrelevant, because an absolute loss limit to achieve power supply efficiency goals becomes dominant.

For the average situation with natural convection cooling, a crude "rule of thumb" can be used:

$$R_E = \frac{800°\text{C - cm}^2\text{ / Watt}}{\text{[illegible]}^2} \quad °\text{C / Watt}$$

Where A_S is the total surface area of the transformer, excluding the mounting surface. Calculating A_S is time-consuming, but another rule of thumb simplifies this, as well. For a given class of cores, such as E-E cores in the ETD or EC series, the relative proportions are quite similar for all core sizes. Thus for all cores in the ETD or EC series, the usable surface area, A_S, is approximately 22 times the winding window area, A_W. Combining this with the equation above enables the window area, A_W, from the core data sheet, to be used to directly calculate the external thermal resistance:

$$R_E = \frac{36}{A_W \text{ in cm}^2} \quad °\text{C / Watt}$$

For pot cores or PQ cores, window areas are proportionately smaller, and not as consistent. A_S/A_W may range from 25 to 50, so that R_E may range from $16/A_W$ to $32/A_W$ °C/W.

Experience is a great help in minimizing and crudely quantifying thermal resistance. In the final analysis, an operational check should be conducted with a thermocouple cemented at the hot spot near the middle of the centerpost, with the transformer mounted in a power supply prototype or mockup.

Worst Case Losses

Transformer losses should be examined under worst-case conditions that the power supply is expected to operate over long periods of time, not under transient conditions.

Transformer losses can be put into three major categories: core hysteresis losses, core eddy current losses, and winding losses.

Core hysteresis losses are a function of flux swing and frequency. In all buck-derived applications under steady-state conditions, $V_{IN} \bullet D = n \bullet V_O$. Under fixed frequency operation, volt-seconds and therefore flux swing are constant. *Hysteresis loss is therefore constant*, regardless of changes in V_{IN} or load current.

Core eddy current loss, on the other hand, is really I^2R loss in the core material. If V_{IN} doubles, Peak I^2R loss quadruples, but since D is halved, average I^2R loss doubles. Thus core eddy current loss is proportional to V_{IN}. *Worst case is at high V_{IN}.*

Winding losses: In buck-derived regulators, peak secondary current equals load current and peak primary current equals load current divided by the turns ratio:

$$I_{Spk} = I_L \;\; ; \;\; I_{Ppk} = I_L / n$$

Peak currents are independent of V_{IN}. But at constant peak currents (constant load), *rms current squared (and I^2R loss) is proportional to duty cycle D and inversely proportional to V_{IN}.*. (With constant peak current, high order harmonics depend mostly on switching transitions and do not vary significantly with D.)

In buck-derived regulators, winding loss is always greatest at low V_{IN}.

Ferrite cores: In most ferrite materials used in SMPS applications, hysteresis losses dominate up to 200-300kHz. At higher frequencies, eddy current losses take over, because they tend to vary with frequency squared (for the same flux swing and waveshape).

Thus, *at frequencies up to 200-300kHz, worst case is at low V_{IN} and full load* because of high winding losses. Once core eddy current losses become significant, they rise rapidly with frequency, especially at high V_{IN}. (The increase in eddy current loss with high V_{IN}, small D, is not shown in core manufacturer's loss curves because they assume sinusoidal waveforms.) Winding losses also rise with frequency, especially at low V_{IN}. To maintain a reasonable R_{AC}/R_{DC}, Litz wire with more strands of finer wire must be used, raising R_{DC} because increased insulation and voids reduce the copper area. Thus, *at frequencies where core eddy current losses dominate, core loss worst case is at high V_{IN}, full load. Winding loss worst case is always at low V_{IN}, full load.*.

Laminated metal alloy and powdered metal cores: Core eddy current losses dominate, hence worst case is at high V_{IN}, full load. Winding losses are worst case at low V_{IN}, full load.

Balancing Core and Winding Losses

At SMPS operating frequencies, when the core is usually loss-limited, not saturation limited, total losses are at a broad minimum when core losses are approximately equal to or a little less than winding losses. Likewise, winding losses are at a minimum and well distributed by making the rms current density approximately equal in all windings. With a bridge or half-bridge primary, which has good winding utilization, and center-tapped secondaries which have poor utilization, rms current densities will be approximately equalized when the primary conductor cross-section area is 40% and the secondaries 60% of the available area. In most other cases, primary and secondary conductor areas should be 50%/50%, including: Forward converter (single-ended primary/secondary SE/SE), C.T. primary/C.T. secondary, bridge-half bridge primary/bridge secondary.

The above allocations can be impossible to achieve because the number of turns in each winding must be an integral number. In a low voltage secondary, 1.5 turns may be required for optimum balance between core and winding losses. With one turn, the flux swing and core loss may be much too large; with two turns the winding loss becomes too great. At either extreme, it may be impossible to meet temperature rise or absolute loss limits. A larger core may be required to resolve this problem.

Window Utilization

This subject is discussed extensively in Section 3. As a reminder:

- Safety isolation requirements impose minimum dimensional limits for creepage and insulation thickness which can waste a high percentage of window area, especially in a small transformer. A bobbin also reduces the area available for windings.

  Triple insulated wire satisfies the insulation thickness requirement and eliminates the creepage requirement. It is worth considering, especially for small transformers where creepage distances take up a large percentage of window area.
- In the reduced window area that is available for the windings, much of the actual winding area is taken up by voids between round wires and by wire insulation. In a winding consisting of many turns of single, round, insulated wires, only 70 - 75% of the area available for that winding is likely to be conductor metal -- "copper". With Litz wire, the copper area is reduced further. For every level of twisting , an additional 0.75 factor (approximate) applies. For example, with Litz wire 7 strands of 7 strands (49 total wires), the copper area would be .75•.75•.75 = 42% of the area available for that winding. On the other hand, a winding consisting of layers (turns) of copper foil or strap, there are no voids, only the insulation between turns. Winding area utilization could be as much as 80 - 90% copper area.

Topology

The choice of circuit topology obviously has great impact on the transformer design, but a detailed discussion is beyond the scope of this topic.

There is a great deal of overlap in topology usage. Flyback circuits (flyback transformers are covered in Section 5) are used primarily at power levels in the range of 0 to 150 Watts, Forward converters in the range of 50 to 500 Watts, half-bridge from 100 to 1000 Watts, and full bridge usually over 500 Watts.

Full bridge and half-bridge topologies with full bridge secondaries have the best transformer efficiency because the core and the windings are fully utilized. With center-tapped secondaries, winding utilization and efficiency are reduced. With center-tapped primary *and* secondaries, winding utilization and efficiency are further reduced. All of the push-pull topologies have the further advantage that for a given switching frequency, giving the same output ripple filtering and closed loop capability, the frequency at which the transformer core and windings operate is halved, reducing core and ac winding losses.

Forward converter transformers have the poorest utilization and efficiency because neither the core nor the windings are used during the lengthy core reset interval.

Frequency

There are several meanings to the term "frequency" in switching power supply applications, and it is easy for confusion to arise.

In this paper, "switching frequency", f_S, is defined as the frequency at which switch drive pulses are generated. It is the frequency seen by the output filter, the frequency of the output ripple and input ripple current, and is an important concept in control loop design. In a single-ended power circuit such as the forward converter, the power switch, the transformer, and the output rectifier all operate at the switching frequency and there is no confusion. The transformer frequency and the switching frequency are the same.

"Clock frequency" is the frequency of clock pulses generated in the control IC. Usually, the switching frequency is the same as the clock frequency, but not always. Occasionally, the control IC may divide the clock frequency to obtain a lower switching frequency. It is not unusual for a push-pull control IC to be used in a single-ended forward converter application, where only one of the two switch drivers is used, to guarantee 50% max. duty cycle. In this case the switching frequency is half the clock frequency.

Confusion often arises with push-pull topologies. Think of the push-pull power circuit as a 2:1 frequency divider, with the transformer and the individual switches and individual rectifiers operating at a "transformer frequency", f_T, which is one-half of the switching frequency. Collectively, the switches and rectifiers operate at the switching frequency, but the transformer operates at the transformer frequency. Some designers define "switching frequency" as the frequency that the individual switch and the transformer operate at, but this requires redefining the term "switching frequency" when dealing with output ripple and in control loop design.

Duty Cycle

Duty cycle, D, is defined as the amount of time the power switch is on in relation to the switching period: $D = t_{ON}/T_S$.

In a single-ended forward converter, this is clearly understood, but in a push-pull circuit, ambiguity often arises. For example, in a half-bridge circuit operating at minimum V_{IN}, the duty cycle is likely to be in the vicinity of 90% (D = 0.9). The transformer is delivering power to the output 90% of the time, there is a voltage pulse applied to the filter input 90% of the time, etc. But individual power switches and individual rectifiers, which conduct only during alternate switching periods, can be said to operate at a duty cycle of 45%. That is true, but it is better tothink of them as operating at D/2, retaining a consistent definition of D throughout the power supply design.

Maximum Duty Cycle

In normal steady-state operation of a buck-derived regulator, $V_{IN} \bullet D$ is constant. The control loop changes duty cycle D inversely proportional to V_{IN} to maintain a constant output voltage, V_O. ($V_{IN}D = n \bullet V_O'$), where n is the turns ratio N_P/N_S, and V_O' equals output voltage V_O plus diode forward voltage drop at full load.

At a fixed switching frequency and *with normal steady-state operation, the volt-seconds applied to the transformer windings are constant,* independent of line voltage or load current.

$$V_{IN}t_{on} = \frac{V_{IN}D}{f_S} = \frac{nV_O'}{f_S}$$

The maximum duty cycle, Dmax, associated with minimum V_{IN} in normal steady-state operation, is limited by a variety of considerations:

In a forward converter, a substantial portion of each switching period must be allowed for core reset. If the voltage backswing during reset is clamped to V_{IN}, the duty cycle must be limited to less than 50% because the time required for reset equals the switch ON time.

In a push-pull converter (bridge, half-bridge, PPCT) duty cycle can approach 100% at the switching frequency (always think of D at the switching frequency, not the transformer frequency). However, it may be necessary to limit D to less than 90% to allow a current transformer to self-reset.

Often the control IC limits the duty cycle for several reasons including allowing time for delays in turning off the switch.

At low V_{IN}, if normal Dmax is right at the duty cycle limit, the regulator has no reserve volt-second capability and cannot respond rapidly to a sudden load increase occuring when V_{IN} is low. It may be desirable to make the "normal" Dmax less than the absolute limit, Dlim, to provide a little headroom in this situation.

A potentially serious problem needs to be considered: During initial start-up of the power supply, or following a sudden large increase in load current which temporarily pulls down Vout, the control loop calls for full current, pushing the duty cycle to its absolute maximum limit, Dlim. The output filter inductor limits the current rate of rise, so that for several switching frequency periods, the duty cycle is at the limit, Dlim. During the transient event described above, Dlim could occur when V_{IN} is maximum. Thus, the volt-seconds applied to the transformer windings could be several times larger than normal:

Limit $V_{IN}D = V_{IN}$maxDlim
Normal $V_{IN}D = V_{IN}$minDmax

The flux swing, also several times greater than normal, could saturate the core. (The increased core loss is not a problem–it is only temporary.)

This may not be a problem if the ratio limit/normal V_{IND} is small and/or if the normal flux density swing, limited by core loss, is a small fraction of Bsat (Bsat - Br for a forward converter). For example, if limit/normal V_{IND} is 3:1, and if normal ΔB is 0.08T, then with Bsat greater than 0.24T, there is no problem.

If this problem exists, soft-start circuitry can eliminate it during start-up, but soft-start has no effect when the load increases rapidly. A few IC control circuits have volt-second limiting capability, but the vast majority do not. The soft saturation characteristic of power ferrite material may be forgiving enough to allow the core to saturate, with the absolute current limit providing protection, but with sharp-saturation core materials, this is a likely disaster. If all else fails, the normal flux swing must be reduced to the point where the abnormal flux swing does not reach saturation.

Restrictions on Number of Turns

Choices regarding the number of turns and turns ratios are often severely limited by low voltage secondaries. For a 5 Volt output the alternatives might be a 1-turn or a 2-turn secondary–a 2 to 1 step in the number of turns in every winding. For the same size core and window, this doubles the current density in the windings and accordingly increases the loss.

Choices may be further restricted when there are multiple low voltage secondaries. For example, a 2.5 to 1 turns ratio may be desirable between a 12 Volt and a 5 Volt output. This is easily accomplished with a 2-turn 5V secondary and a 5-turn 12V winding. But if the 5V secondary has only 1 turn, the only choice for the 12V secondary is 3 turns, which may result in excessive linear post-regulator loss. This problem could be handled by the use of fractional turns -- see reference (R6).

There are no hard and fast rules to follow in establishing the optimum turns for each winding, but there is some general guidance. First, define the ideal turns *ratios* between windings that will achieve the desired output voltages with the normal V_{IND} established earlier. Later, when a specific core has been tentatively selected, the turns ratios will translate into specific turns, but these are not likely to be the integral numbers required in practice. It then becomes a juggling act, testing several approaches, before reaching the best compromise with integral turns. The lowest voltage secondary usually dominates this process, because with small numbers the jumps between integral turns are a larger percentage. Especially if the lowest voltage output has the greatest load power, which is often the case, the lowest voltage secondary is rounded up or down to the nearest integral. Rounding down will increase core loss, rounding up will increase winding loss. If the increased loss is unacceptable, a different core must be used that will require less adjustment to reach an integral number of turns. The low voltage output is usually regulated by the main control loop.

Higher voltage secondaries can be rounded up to the next integral with less difficulty because they have more turns. However, it is unlikely that accuracy or load regulation will be acceptable, requiring linear or switched post-regulation.

Since the primary is usually higher voltage, the primary turns can usually be set to achieve the desired turns ratio without difficulty.

Once the turns have been established, the initial calculations must be redefined.

Flux Walking

Faraday's Law states that the flux through a winding is equal to the integral volt-seconds per turn. This requires that the voltage across any winding of any magnetic device must average zero over a period of time. The smallest dc voltage component in an applied ac waveform will slowly but inevitably "walk" the flux into saturation.

In a low frequency mains transformer, the resistance of the primary winding is usually sufficient to control this problem. As a small dc voltage component pushes the flux slowly toward saturation, the magnetizing current becomes asymmetrical. The increasing dc component of the magnetizing current causes an IR drop in the winding which eventually cancels the dc voltage component of the drive waveform, hopefully well short of saturation.

In a high frequency switchmode power supply, a push-pull driver will theoretically apply equal and opposite volt-seconds to the windings during alternate switching periods, thus “resetting” the core (bringing the flux and the magnetizing current back to its starting point). But there are usually small volt-second asymmetries in the driving waveform due to inequalities in MOSFET R_{DSon} or switching speeds. The resulting small dc component will cause the flux to “walk”. The high frequency transformer, with relatively few primary turns, has extremely low dc resistance, and the IR drop from the dc magnetizing current component is usually not sufficient to cancel the volt-second asymmetry until the core reaches saturation.

Flux walking is not a problem with the forward converter. When the switch turns off, the transformer magnetizing current causes the voltage to backswing, usually into a clamp. The reverse voltage causes the magnetizing current to decrease back to zero, from whence it started. The reverse volt-seconds will exactly equal the volt-seconds when the switch was ON. Thus the forward converter automatically resets itself (assuming sufficient reset time is allowed, by limiting the maximum duty cycle).

The flux walking problem is a serious concern with any push-pull topology (bridge, half-bridge or push-pull CT), when using voltage mode control..

One solution is to put a *small* gap in series with the core. This will raise the magnetizing current so that the IR drop in the circuit resistances will be able to offset the dc asymmetry in the drive waveform. But the increased magnetizing current represents increased energy in the mutual inductance which usually ends up in a snubber or clamp, increasing circuit losses.

A more elegant solution to the asymmetry problem is an automatic benefit of using current mode control (peak or average CMC). As the dc flux starts to walk in one direction due to volt-second drive asymmetry, the peak magnetizing current becomes progressively asymmetrical in alternate switching periods. However, current mode control senses current and turns off the switches at the same peak current level in each switching period, so that ON times are alternately lengthened and shortened. The initial volt-second asymmetry is thereby corrected, peak magnetizing currents are approximately equal in both directions, and flux walking is minimized.

However, with the half-bridge topology this creates a new problem. When current mode control corrects the volt-second inequality by shortening and lengthening alternate pulse widths, an ampere-second (charge) inequality is created in alternate switching periods. This is of no consequence in full bridge or push-pull center-tap circuits, but in the half-bridge, the charge inequality causes the capacitor divider voltage to walk toward the positive or negative rail. As the capacitor divider voltage moves away from the mid-point, the volt-second unbalance is made worse, resulting in further pulse width correction by the current mode control. A runaway situation exists, and the voltage will walk (or run) to one of the rails. This problem is corrected by adding a pair of diodes and a low-power winding to the transformer, as detailed in the Unitrode Applications Handbook.

Core Selection: Material

Select a core material appropriate for the desired *transformer* frequency.

With power ferrites, higher frequency materials have higher resistivity, hence lower eddy current losses. However, the permeability is generally lower, resulting in greater magnetizing current, which must be dealt with in snubbers and clamps.

With metal alloy cores, the higher frequency materials have higher resistivity and require very thin laminations. Although saturation flux density is usually very much greater than with ferrite materials, this is usually irrelevant because flux swing is severely limited by eddy current losses.

Ferrite is the best choice in transformer applications except for mechanical ruggedness.

Core Selection: Shape

The window configuration is extremely important. The window should be as wide as possible to maximize winding breadth and minimize the number of layers. This results in minimized Rac and leakage inductance. Also, with a wide window, the fixed creepage allowance dimension has less impact. With a wider window, less winding height is required, and the window area can be better utilized.

Pot cores and PQ cores have small window area in relation to core size, and the window shape is almost square. The creepage allowance wastes a large fraction of the window area, and the winding breadth is far from optimum. These cores are not as well suited for high frequency SMPS applications. One advantage of pot cores and PQ cores is that they provide better magnetic shielding than E-E cores, reducing EMI propagation.

EC, ETD, LP cores are all E-E core shapes. They have large window area in relation to core size, and the window has the desirable wide configuration.

Toroidal cores, properly wound, must have all windings distributed uniformly around the entire core. Thus the winding breadth is essentially the circumference of the core, resulting in the lowest possible leakage inductance and minimizing the number of winding layers. There is no creepage allowance because there is no end to the windings. (But there is a problem bringing the leads out.) Stray magnetic flux and EMI propagation are also very low.

The big problem with toroidal cores is the winding difficulty, especially with the shapes and gauge of conductors used in SMPS transformers. How can a 1-turn secondary be spread around the entire toroid? Automatic winding is virtually impossible. For this reason, toroidal shapes are seldom used in SMPS transformers.

Planar cores with their low profile are becoming more popular as SMPS frequencies progressively increase. Planar cores introduce a new set of unique problems which are beyond the scope of this discussion. Be assured that Faraday's and Ampere's Laws still apply, but in a planar core, flux density and field intensity change considerably throughout the important regions, making calculation much more difficult.

Core Selection: Size

A novice in the art of transformer design usually needs some guidance in making an initial estimate of the core size appropriate for the application requirements. One widely used method, with many variations, is based on the core Area Product, obtained by multiplying the core magnetic cross-section area by the window area available for the winding.

There are many variables involved in estimating the appropriate core size. Core power handling capability does not scale linearly with area product or with core volume. A larger transformer must operate at a lower power density because heat dissipating surface area increases less than heat-producing volume. The thermal environment is difficult to evaluate accurately, whether by forced air or natural convection.

Some core manufacturers no longer provide area product information on their data sheets, often substituting their own methodology to make an initial core size choice for various applications.

The following formula provides a crude indication of the area product required:

$$AP = A_W A_E = \left(\frac{P_O}{K \, \Delta B \, f_T} \right)^{4/3} \quad cm^4$$

where:

P_O	= Power Output
ΔB	= Flux density swing, Tesla
f_T	= *Transformer* operating frequency
K	= .014 (Forward converter,PPCT)
	= .017 (Bridge, half bridge)

This formula is based on current density of $420A/cm^2$ in the windings, and assumes a window utilization of 40% copper. At low frequencies, the flux swing is limited by saturation, but above 50kHz (ferrite), ΔB is usually limited by core losses. Use the ΔB value that results in a core loss of $100mW/cm^3$ (2 times the "flux density" given in the core loss curves).

These initial estimates of core size are not very accurate, but they do reduce the number of trial solutions that might otherwise be required. In the final analysis, the validity of the design should be checked with a prototype transformer operated in the circuit and the environment of the application, with the hot spot temperature rise measured by means of a thermocouple cemented to the center of the centerpost.

Transformer Design Cookbook

The steps for designing a power transformer for SwitchMode Power Supplies is outlined below. A typical example is carried through to illustrate the process. There are many approaches to transformer design. The approach presented here appears the most logical and straightforward to the author.

It may be worthwhile to use software such as "Magnetic Designer" from Intusoft[2] for the initial design, using the approach defined herein for verification and tune-up. The author has not evaluated "Magnetic Designer" sufficiently to make an unqualified endorsement, but it should certainly make a good starting point and take a great deal of drudgery out of the process. It has the advantage of including an extensive core database.

Initial Preparation

The first few steps in this process define application parameters that should not change, regardless of subsequent iterations in the selection of a specific core type and size.

If the results are not acceptable, start over from the very beginning, if that seems appropriate. Great difficulty in achieving an acceptable forward converter transformer design may be a subtle message that a half-bridge topology is perhaps a better choice.

Step 1. Define the power supply parameters pertaining to the transformer design:

V_{IN} Range:	100 - 190 V
Output 1:	5 V, 50 A
Output 2:	none
Circuit Topology:	Forward Converter
Switching Freq, f_S:	200 kHz
Transformer Freq, f_T:	200 kHz
Max Loss (absolute):	2.5 W
Max °C Rise:	40°C
Cooling Method:	Natural Convection

Step 2. Define absolute duty cycle limit Dlim, tentative normal Dmax at low V_{IN} (to provide headroom for dynamic response), and normal $V_{IN}D$:

Absolute Limit, Dlim:	0.47
Normal Dmax:	0.42
Normal V_{IN}•D :	V_{IN}min•Dmax = 42 V
V_{IN}maxDlim:	89.3 V

Step 3. Calculate output voltages plus diode and secondary IR drops at full load:

Vo1': 5.0 + 0.4 = 5.4 Volts
Vo2': n/a

Step 4. Calculate desired turns ratios: P-S1; S1-S2, etc. Remember that choices with low voltage secondaries will probably be limited.

$n = N_P/N_{S1} = V_{IN}D/V_{O1}'$: 42/5.4 = 7.8
Possible choices: 8:1 ; 7:1 ; 15:2

Core Selection

Step 5: Tentatively select core material, shape and tentative size, using guidance from the manufacturer's data sheet or using the area product formula given previously in this paper. Will a bobbin be used?:

Core Material: Ferrite, Magnetics Type P
Core type, Family: ETD
Core Size: 34mm -- ETD34

Step 6: For the specific core selected, note:
Effective core Area, Volume, Path Length. (cm)

A_e: 0.97 cm^2
V_e: 7.64 cm^3
ℓ_e: 7.9 cm

Window Area, Breadth, Height, Mean Length per Turn (' indicates net with bobbin, creepage).

A_w / A_w': 1.89 / 1.23 cm^2
bw / bw': 2.36 / 1.5 cm
hw / hw': 0.775 / 0.6 cm
MLT: 5.8 / 6.1 cm

Define R_T and Loss Limit

Step 7: Obtain thermal resistance from data sheet or calculate from window area (not bobbin area) from formula for EC and ETD series:

$$R_E = \frac{800}{22 \bullet A_W \, \text{in cm}^2} = \frac{36}{1.89} = 19 \text{ °C/Watt}$$

Calculate loss limit based on max. temperature rise:

Plim = °Crise/R_T: = 40/19 = 2.1 Watts

The 2.1W limit applies, since it is less than the absolute limit from Step 1. Tentatively apportion half to core loss, half to winding loss.

PClim: 1 Watt

Pwlim: 1.1 Watt

Step 8: Loss Limited Flux Swing

Calculate max. core loss per cm^3

PClim/Ve = 1/7.64 = 131 mw/cm^3 (= kW/m^3)

Using this core loss value, enter the core loss curve for the P material selected. At the *transformer* frequency, find "flux density" (actually peak flux density). Double it to obtain the loss-limited peak-peak flux density swing, ΔB:

At 131 mw/cm^3 and 200kHz:

ΔB = 2•800 Gauss = 1600G = 0.16 Tesla

Normal $\Delta\phi = \Delta B \cdot A_e$

Step 9: Using Faraday's Law, calculate the number of secondary turns:

$$\int E_{S1}\,dt = Vpk_{S1}t_{ON} = V_{O1}'\cdot T_S$$

$$N_{S1} = \int E_{S1}dt/\Delta\phi = V_{O1}'\cdot T_S/\Delta\phi$$

$$N_{S1} = \frac{V_{O1}'T_S}{\Delta B \times A_E} = \frac{5.4 \times 5 \times 10^{-6}}{.16 \times .97 \times 10^{-4}} = 1.74 \text{ Turns}$$

Rounding down to 1 turn will greatly increase the volts/turn, flux swing and core losses. Rounding up to 2 turns reduces core losses but increases winding loss. Since the result above is much closer to 2 turns, this will be adopted.

Step 10: Recalculate flux swing and core loss at 2 turns:

$$\Delta B(\text{2 turns}) = 0.16\text{T}\frac{1.74\text{turns}}{2\text{ turns}} = 0.14\text{Tesla}$$

From the core loss curves, loss at 0.14T/2 (700 Gauss) is 110mw/cm^3 x 7.64cm^3

Core loss = 0.84 W

Step 11: Finalize the choice of primary turns. A larger turns ratio results in lower peak current, larger D (less reserve), and more copper loss. From the possibilities defined in Step 4, trial solutions show the best choice to be N_P = 15 turns (7.5:1 turns ratio).

Recalculate normal V_{IND} and flux swing under worst case V_{IN}maxDlim conditions:

V_{IND} = nVo' = 7.5•5.4 = 40.5 V

ΔBlim = 0.14T•89.3/40.5 = 0.31T -- OK

Step 12: Define the winding structure.

An interleaved structure will be used, as shown in Figure 4-1, to minimize leakage inductance and winding losses.

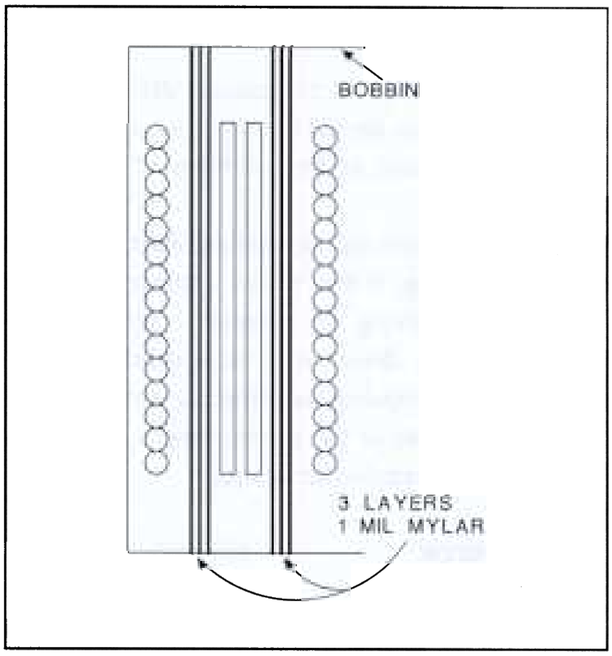

Figure 4-1

The interleaved structure results in two winding sections. Primary windings of 15 turns in each section are connected in parallel. Primary current divides equally in the two paralleled windings because this results in the lowest energy transfer. Secondary windings of 1 turn copper foil in each section are connected in series, resulting in a 2-turn secondary. With only one turn in each section, the secondary windings can be much thicker than DPEN to minimize dc resistance without increasing the ac resistance.

Step 12: Calculate DPEN at 200 kHz:

$$Dpen = 7.6/\sqrt{f} = 7.6/\sqrt{200{,}000} = .017 \text{ cm}$$

Step 13: Calculate dc and rms ac currents in each winding at VINmin and Dmax. (Ref. Section 3):

$$ISdc = 50A \cdot Dmax = 50 \cdot 0.405 = 20.25A$$

$$ISac = ISdc((1\text{-}D)/D)^{1/2} = 24.5A$$

$$IPdc = ISdc/n = 20.25/7.5 = 2.7A$$

$$IPac = ISac/n = 24.5/7.5 = 3.27A$$

Primary current in each of the two paralleled sections is one-half the total primary current: 1.35Adc and 1.65Aac.

Step 14: Define the primary winding:

One layer of 15 turns spread across the available winding breadth of 1.3cm allows a maximum insulated wire diameter of 0.87mm. AWG 21 – 0.72mm copper will be used.

From Ref R2, pg 9, the effective layer thickness equals $0.83 \cdot dia(dia/spacing)^{1/2}$.

$Q = $ (layer thickness)/DPEN

$Q = 0.83 \cdot .072(.072/.087)^{1/2}/.017 = 3.19$

From Dowell's curves, Rac/Rdc for 1 layer is 3.1. This will result in unacceptable ac losses.

A Litz wire consisting of 100 strands #42 wire has a diameter of 0.81mm and a resistance of 0.545mΩ/cm.

The dc resistance of the single layer is:

$$Rdc = \Omega/cm \cdot MLT \cdot Ns = .00055 \cdot 6.1 \cdot 15 = .05\Omega$$

Multiplying by $(1.35Adc)^2$, dc power loss is .091W in each section, for a total primary dc loss of 0.18W.

The diameter of each #42 wire is .064mm, but there are effectively ten layers of fine wire in the single layer of Litz wire. This is because the 100 strands are roughly equivalent to a 10 x 10 array, thus ten wires deep. Q is approximately 1/10 the value for solid wire, or 0.3, resulting in Rac/Rdc of 1.2. Thus, $Rac = Rdc \cdot 1.2$, or .06Ω.

Multiplying by 1.65A squared, the ac loss is 0.16W in each section, for a total primary ac loss of 0.32W. Adding the 0.18W dc loss,

Total primary power loss = 0.5 Watts.

Step 15: Define the secondary winding.

The secondary consists of two turns (two layers) of copper strip or foil, 1.3cm wide (full available winding breadth), and 0.13cm thick. There is one secondary layer in each of the two sections of the interleaved winding structure. This permits the thickness of the copper strip to be much greater than DPEN to minimize dc losses, without increasing ac losses. This is because ac current flows only on the outer side of each turn. As the conductor is made thicker, Rac/Rdc becomes larger, but Rdc decreases and Rac remains the same.

With a solid copper secondary, the layer thickness is the same as the conductor thickness, 0.1cm.

$Q = $ Layer thickness/DPEN $= 0.13/.017 = 7.6$

$Rac/Rdc = 7.5$

This will be acceptable because the dc resistance is very low.

$Rdc = \rho \cdot MLT \cdot Ns/(bw' h)$

$Rdc = 2.3 \cdot 10^{-6} \cdot 6.1 \cdot 2/(1.3 \cdot 0.13) = 166\mu\Omega$

$Pdc = 166\mu\Omega \cdot 20.25^2 = .068 \text{ W}$

$Pac = Rdc \cdot Rac/Rdc \cdot Iac^2 = 166\mu\Omega \cdot 7.5 \cdot 24.5^2$

$Pac = 0.75 \text{ W}$

Total secondary loss:

.068W + 0.75W = 0.82 W

Total copper loss:

0.82W + 0.5W = 1.32 W

Total core plus copper loss:

0.84W + 1.32W = 2.16 Watts

Thus, the total power loss is under the absolute limit of 2.5Watts, but slightly over the 2.1 Watt limit based on the desired max. temperature rise of 40°C.

References:

(R2) "Eddy Current Losses in Transformer Windings and Circuit Wiring," *Unitrode Seminar Manual SEM600*, 1988 (reprinted in the Reference Section at the back of this Manual)

(R4) "The Effects of Leakage Inductance on Switching Power Supply Performance," *Unitrode Seminar Manual SEM100*, 1982 (reprinted in the Reference Section at the back of this Manual)

(R6) "How to Design a Transformer with Fractional Turns," *Unitrode Seminar Manual SEM500*, 1987 (reprinted in the Reference Section at the back of this Manual)

(1) PROXY -- Proximity effect analysis, KO Systems, Chatsworth, CA, 818-341-3864

(2) "Magnetics Designer," Magnetics design software, IntuSoft, San Pedro, CA 310-833-0710

Section 5 Inductor and Flyback Transformer Design

Filter inductors, boost inductors and flyback transformers are all members of the "power inductor" family. They all function by taking energy from the electrical circuit, storing it in a magnetic field, and subsequently returning this energy (minus losses) to the circuit. A flyback transformer is actually a multi-winding coupled inductor, unlike the true transformers discussed in Section 4, wherein energy storage is undesirable.

Application Considerations

Design considerations for this family of inductors vary widely depending on the type of circuit application and such factors as operating frequency and ripple current.

Inductor applications in switching power supplies can be defined as follows (see Fig. 5-1):

- *Single winding inductors:*
  - Output filter inductor (buck-derived)
  - Boost inductor
  - Flyback (buck-boost) inductor
  - Input filter inductor
- *Multiple winding inductors:*
  - Coupled output filter inductor [R5]
  - Flyback transformer

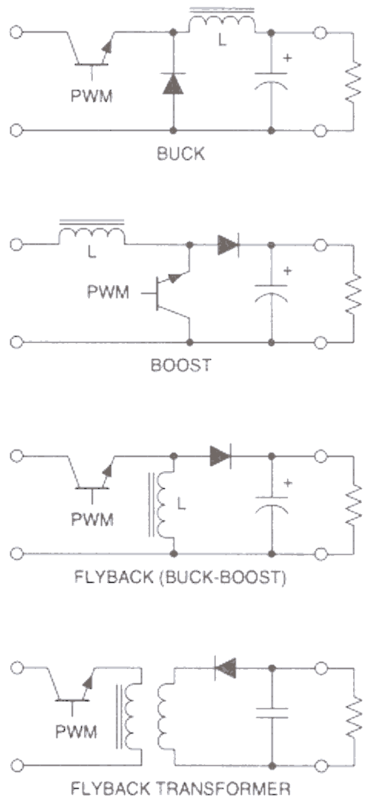

Figure 5-1 Inductor Applications

Inductor design also depends greatly on the inductor current operating mode (Figure 5-2):

- *Discontinuous inductor current mode,* when the instantaneous ampere-turns (totaled in all windings) dwell at zero for a portion of each switching period.
- *Continuous inductor current mode,* in which the total ampere-turns do not dwell at zero (although the current may pass *through* zero).

In the continuous current mode, the ripple current is often small enough that ac winding loss and ac core loss may not be significant, but in the discontinuous mode, ac losses may dominate.

Design limitations: The most important limiting factors in inductor design are (a) temperature rise and efficiency considerations arising from core losses and ac and dc winding losses, and (b) core saturation.

Output filter inductors (buck-derived) --single and multiple windings are seldom operated in the *discontinuous* current mode because of the added burden this places on the output filter capacitor, and because it results in poor cross-regulation in multiple output supplies. Typically operated in the *continuous mode* with peak-peak ripple current much smaller than full load current, ac winding loss is usually not significant compared to dc loss.

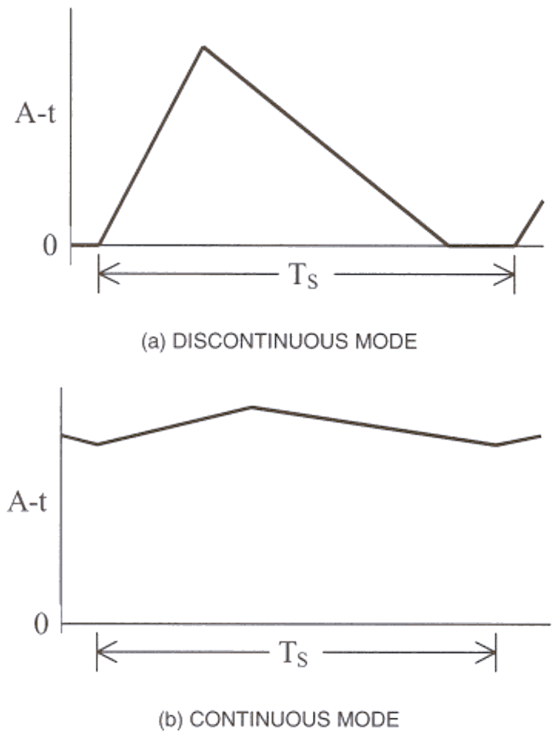

Figure 5-2 Inductor Current Modes

For example, assume full load I_{dc} of 10A, and typical peak-peak triangular ripple current 20% of I_{dc}, or 2A (worst at high Vin). In this example, the worst-case rms ripple current is 0.58A (triangular waveform rms equals $I_{pp}/\sqrt{12}$), and rms ripple current *squared* is only .333, compared with the dc current *squared* of 100. Thus, for the ac I^2R loss to equal the dc loss, the R_{ac}/R_{dc} ratio would have to be as large as 300 (Section 3, Fig. 3-5). This is easily avoided. Therefore, ac winding loss is usually not significant.

Also, the small flux swing associated with small ripple current results in small core loss, with high frequency ferrite core material operating below 250kHz. Core utilization is then limited by saturation (at peak short-circuit current). However, the small flux swing may permit the use of lossier core materials with higher B_{SAT}, such as powdered iron, Kool-mu®, or laminated metal. This may enable reduced cost or size, but core loss then becomes more significant. Also, distributed-gap materials exhibit rounding of the B-H characteristics (Sec. 2, pg. 2-3), resulting in decreasing inductance value as current increases.

Boost and input filter inductors and single winding flyback inductors are often designed to operate in the *continuous mode*. As with the buck-derived filter inductors described previously, inductor design is then usually limited by dc winding losses and core saturation.

However, many boost and flyback applications are designed to operate in the *discontinuous mode*, because the required inductance value is less and the inductor physical size *may* be smaller. But in the discontinuous mode, the inductor current must dwell at zero (by definition) during a portion of each switching period. Therefore, the peak of the triangular current waveform, and thus the peak-to-peak ripple must be *at least twice* the average current, as shown in Fig. 5-2(a). This very large ripple current results in a potentially serious ac winding loss problem. Also, the resulting large flux swing incurs high core loss. Core loss then becomes the limiting factor in core utilization, rather than saturation, and may dictate a larger core size than otherwise expected.

Thus, the circuit designer's choice of operating mode makes a substantial difference in the inductor design approach.

When **flyback transformers** are operated in the continuous inductor current mode, the total ampere-turns of all the windings never dwell at zero (by definition). However, the current in *each winding* of any flyback transformer is *always highly discontinuous*, regardless of inductor current mode. This is because current (ampere-turns) transfers back and forth between primary and secondary(s) at the switching frequency. As shown in Fig. 5-3, the current in each winding alternates from zero to a high peak value, even though the *total* ampere-turns are continuous with small ripple. This results in large ac winding loss, regardless of the operating mode.

However, the core sees the *total* ampere-turn ripple. Thus, core loss behaves in the same manner as with the single winding flyback inductor discussed previously – small core loss when designed and operated in the continuous mode, large core loss in the discontinuous mode.

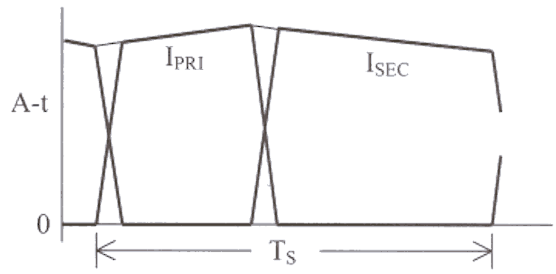

Figure 5-3 – Flyback Transformer Currents

Losses and Temperature Rise

The discussion in Section 4 regarding temperature rise limits, losses and thermal resistance in transformers (pp 4-1,2) is generally applicable to inductors, as well.

Balancing Core and Winding Losses

When inductors are designed for the discontinuous mode, with significant core loss, total loss is at a broad minimum when core and winding losses are approximately equal. But when inductors are designed for the continuous mode, core loss is often negligible, so that the total loss limit can be allocated entirely to the windings.

General Considerations -- Core

Ideal magnetic materials cannot store energy. Practical magnetic materials store very little energy, most of which ends up as loss. In order to store and return energy to the circuit efficiently and with minimal physical size, a small non-magnetic gap is required in series with a high permeability magnetic core material. In ferrite or laminated metal alloy cores, the required gap is physically discrete, but in powdered metal cores, the gap is *distributed* among the metal particles.

Paradoxically, virtually all of the magnetic energy is stored the so-called "non-magnetic" gap(s). The sole purpose of the high permeability core material is to provide an easy, low reluctance flux path to link the energy stored in the gap to the winding, thus efficiently coupling the energy storage location (the gap) to the external circuit.

In performing this critically important function, the magnetic core material introduces problems: (a) core losses caused by the flux swings accompanying the storage and release of energy, and (b) core saturation, where the core material becomes non-magnetic and therefore high reluctance above a certain flux density level. The energy storage capability of a practical gapped core is thus limited either by temperature rise associated with core loss, or by core saturation.

Stray Flux. Another problem that must be faced is stray flux, associated with energy stored in a fringing field outside the gap. Stray flux couples noise and EMI to the external circuit and to the outside world. This stray energy also increases the inductance beyond its intended value by an amount that is difficult to predict.

To minimize stray flux, it is very important that the winding distribution conforms to the gap. When the gap is distributed throughout the core, as in powdered metal cores, the winding(s) should be likewise distributed. Thus, a toroidal core shape should have the windings distributed uniformly around the entire core.

With a discrete gap, used with laminated metal alloy cores or ferrite cores, the winding should be directly over the gap. For example, if a pair of "C" core halves has a gap in one leg and the winding is placed on the opposite (ungapped) leg, as shown in Fig. 5-4a, the entire magnetic force introduced by the winding appears across the two core halves. This results in considerable stray flux propagated external to the device, in addition to the flux through the gap. The energy stored in the external stray field can easily equal the energy stored in the gap, resulting in an inductance value much greater than expected. The external stored energy is difficult to calculate, making the total inductance value unpredictable. Also, the additional flux in the stray field will cause the inductor to saturate prematurely.

However, when the same winding is placed on the gapped core leg, as in Fig. 5-4b, the entire magnetic force introduced by the winding is dropped across the gap directly beneath. The magnetic force across the two core halves is then nearly zero, and there is little external flux. The core then serves its intended purpose of providing an easy (low reluctance) return path for the flux, requiring very little magnetic force to do so, and propagating very little external field.

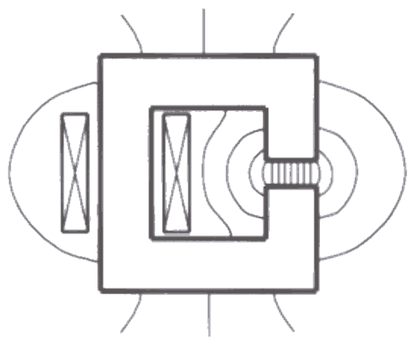

Fig. 5-4a Large External Field

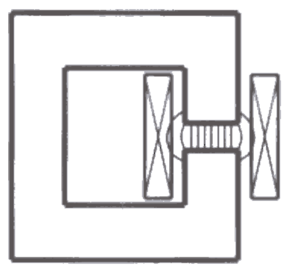

Fig. 5-4b Minimal External Field

Gap area correction: Even when the winding is properly placed directly over a discrete gap, there will be a small but intense fringing field adjacent to the gap, extending outward beyond the boundaries of the core cross-section as shown in Fig 5-4b. Because of this fringing field, the *effective* gap area is larger than the core center-pole area. To avoid what could be a significant error, the inductance calculation must be based upon the effective gap area rather than the actual center-pole area. An empirical approximation is obtained by adding the *length* of the gap to the dimensions of the core center-pole cross-section.[1]

For a core with a rectangular center-pole with cross-section dimensions a and b, the effective gap area, A_g is approximately:

$$A_{cp} = a \times b \quad ; \quad A_g \approx (a + \ell_g) \times (b + \ell_g) \qquad \textbf{(1a)}$$

For a round center-pole with diameter D_{cp}:

$$A_{cp} = \frac{\pi}{4} D_{cp}^2 \quad ; \quad A_g \approx \frac{\pi}{4}\left(D_{cp} + \ell_g\right)^2$$

Resulting in a gap area correction factor:

$$\frac{A_g}{A_{cp}} \approx \left(1 + \frac{\ell_g}{D_{cp}}\right)^2 \qquad \textbf{(1b)}$$

Thus, when ℓ_g equals $0.1D_{cp}$, the area correction factor is 1.21. The gap must be made larger by this same factor to achieve the desired inductance (see Eq. 3a).

The preceding correction factor is helpful when the correction is less than 20%. A more accurate correction requires finite element analysis evaluation, or trial-and-error evaluation.

After the number of turns and the gap length have been calculated according to steps 7 and 8 of the cookbook design procedure presented later in this section, verification is obtained by building a prototype inductor.

If the measured inductance value is too large, *do not* reduce the number of turns, or excessive core loss and/or saturation may result. Instead, increase the gap to reduce the inductance.

If the measured inductance is too small, the number of turns may be increased, but the core will then be under-utilized and winding losses may be excessive. It is best to raise the inductance by decreasing the gap length.

Melted windings: Another serious problem can result from the fringing field adjacent to the gap. Any winding turns positioned close to the gap will likely exist within the high flux density of the fringing field. In applications with large flux swings, huge eddy current losses can occur in those few turns close to the gap. Windings have been known to melt in this vicinity. This problem is most severe with flyback transformers and boost inductors designed for the discontinuous mode, because the flux swings at full load are very large. With filter inductors, or any inductors designed for continuous mode operation, flux swing is much less and the problem is much less severe.

Solutions for devices designed to operate with large flux swing: (1) Don't put winding turns in the immediate vicinity of the gap. Although the winding should be on the center-pole directly over the gap, a non-magnetic, non-conductive spacer could be used to substitute for the turns in the area where the fringing field is strong. (2) Distribute the gap by dividing it into two or three (or more) smaller gaps spaced uniformly along the center-pole leg under the winding. Since the fringing field extends out from the core by a distance proportional to the gap, several small gaps will dramatically reduce the extent of the fringing field. This also results in more accurate inductance calculation. (3) Eliminate the fringing field entirely by using a ferrite core with a powdered metal rod substituted for the ferrite center-leg. This distributes the gap uniformly among the metal particles, directly beneath the entire length of the winding, eliminating the fringing field. While this last method has been used successfully, it is usually not practical because of high cost, and greater ac core losses with metal powder cores.

Gapping all legs: It is tempting to avoid the cost of grinding the gap in the centerleg by merely spacing the two core halves apart, thus placing half the gap in the centerleg and the other half in the combined outer legs. But the outer leg gaps clearly violate the principle that the winding should be placed directly over the gap. A little more than half of the total magnetic force will exist across the centerleg gap, but the remaining force appears across the outer leg gap(s) and thus across the two core halves. This propagates considerable stray flux outside the inductor, radiating EMI, and the inductance value becomes larger and difficult to predict. The result is intermediate between Figures 5-4a and 5-4b.

There is one other benefit of spacing the core halves apart. Because the gap is divided, the smaller centerleg gap length reduces the fringing field and thus reduces the eddy current problem in the winding close to the gap.

A trick which greatly reduces the external stray flux in this situation is to place an external shorted turn around the entire outer periphery of the inductor. The shorted turn is made of wide copper strip placed co-axial with the inductor winding, encircling the entire outer surface of the inductor, outside the windings and outside the outer core legs, and closely conforming to the external shape. Any stray flux that escapes to the outside world will link to this external shorted turn, inducing in it a current which creates a magnetic field in opposition to the stray flux.

Core Selection: Material

Select a core material appropriate for the desired frequency and inductor current mode.

Ferrite is usually the best choice for inductors designed to operate in the *discontinuous* mode at frequencies above 50kHz, when core loss associated with large flux swing limits core utilization.

However, in the continuous mode, with small ripple current and small flux swing, ferrite cores will often be limited by saturation. In this case, lossier core materials with greater saturation flux density, such as powdered iron, Kool-mu®, Permalloy powder, or even gapped laminated metal cores may enable reduced cost or size. But the rounded B-H characteristic of powdered metal cores can result in the perhaps unintended characteristic of a "swinging choke," whose inductance decreases at higher current levels.

Core Selection: Shape

The core shape and window configuration is not critically important for inductors designed to operate in the continuous mode, because ac winding loss is usually very small.

But for inductors designed for discontinuous mode operation, and especially for flyback transformers, the window configuration is extremely important. The window should be as wide as possible to maximize winding breadth and minimize the number of layers. This minimizes ac winding resistance. For a flyback transformer, the wide window also minimizes leakage inductance, and the required creepage distance when line isolation is required has less impact. With a wider window, less winding height is required, and the window area utilization is usually better.

As discussed in Section 4, pot cores and PQ cores have small window area in relation to core size, and the window shape is not well suited for flyback transformers or discontinuous mode inductors.

EC, ETD, LP cores are all E-E core shapes, with large, wide windows which make them excellent choices with ferrite materials. These core shapes lend themselves to spiraled windings of wide copper strip, especially with inductors operated in the continuous mode, where ac winding losses are small.

Toroidal powdered metal cores, with windings distributed uniformly around the entire core, can be used in any inductor or flyback transformer application. Stray magnetic flux and EMI propagation is very low.

But a gapped ferrite toroidal core is a very bad choice. Windings distributed around the toroid will not conform to discrete gaps, resulting in large stray fields, radiated EMI, and inductance values that cannot be calculated.

Optimum core utilization

The smallest size and lowest cost inductor is achieved by fully utilizing the core. In a specific application, optimum core utilization is associated with a specific optimum gap length (resulting in a specific effective permeability μ_e for cores with distributed gaps). The same core in a different application or at a different frequency may have a different optimum gap length.

The optimum gap length results in the core operating at maximum flux density (limited either by saturation or by core loss), and also at maximum current density in the windings (limited by winding loss). This is the best possible utilization of the core, resulting in the smallest size. The inductor design approach should therefore seek to achieve this optimum gap length (or optimum μ_e for a core with distributed gap).

Figure 5-5 shows the characteristic of a core with optimum gap, limited by core saturation and by max. current density in the windings. The area between the characteristic and the vertical axis indicates the energy storage capability. Any other slope (different gap size) results in less energy storage capability

Core Selection: Size

The discussion of core size for transformers in Section 4-8 is mostly relevant to inductors as well. The following Area Product formulae are intended to help provide a rough initial estimate of core size for inductor applications.

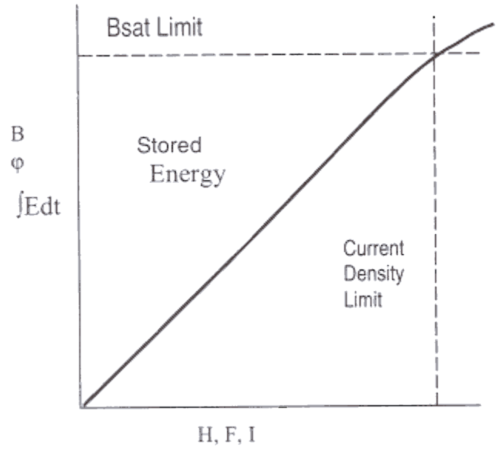

Fig 5-5 Optimum Core Utilization

When core loss is not severe, so that flux swing is limited by core saturation:

$$AP = A_W A_E = \left(\frac{L\, I_{SCpk}}{B_{MAX}} \cdot \frac{I_{FL}}{K_1} \right)^{4/3} \quad \text{cm}^4 \qquad \textbf{(2a)}$$

When flux swing is limited by core loss:

$$AP = A_W A_E = \left(\frac{L\, \Delta I}{\Delta B_{max}} \cdot \frac{I_{FL}}{K_2} \right)^{4/3} \quad \text{cm}^4 \qquad \textbf{(2b)}$$

where:

- L = inductance, Henrys
- I_{SCpk} = max pk short-circuit current, A
- B_{MAX} = saturation limited flux density, T
- ΔI = current swing, Amps (primary)
- ΔB_{max} = max flux density swing, Tesla
- I_{FL} = rms current, full load (primary)
- K_1, K_2 = $J_{MAX} K_{PRI} \times 10^{-4}$

where:

- J_{MAX} = max. current density, A/cm^2
- K_{PRI} = primary copper area/window area
- 10^{-4} = converts dimensions from meters to cm

Application	K_{PRI}	K_1	K_2
Inductor, single winding	0.7	.03	.021
Filter Inductor, multiple winding	.65	.027	.019
Flyback transformer – non-isolated	0.3	.013	.009
Flyback transformer – with isolation	0.2	.0085	.006

For a single winding inductor, the term "primary" above refers to the entire winding.

K_{PRI} represents the utilization of the window containing the winding. For a single winding inductor, K_{PRI} is the ratio of the total copper area to the window area, A_W. For a flyback transformer, K_{PRI} is the ratio of the *primary* winding copper cross-section area to the total window area.

The saturation-limited formula assumes winding losses are much more significant than core losses. K1 is based on the windings operating at a current density of $420A/cm^2$, a commonly used "rule of thumb" for natural convection cooling.

In the core loss limited formula, core and winding losses are assumed to be approximately equal. Therefore, the winding losses are halved by reducing the current density to $297A/cm^2$ (470 x 0.707). Thus, K2 equals 0.707·K1.

In either formula, it is assumed that appropriate techniques are used to limit the increase in winding losses due to high frequency skin effect to less than 1/3 of the total winding losses.

Forced air cooling permits higher losses (but with reduced efficiency). K values become larger, resulting in a smaller core area product.

The 4/3 power shown in both area product formulae accounts for the fact that as core size increases, the volume of the core and windings (where losses are generated) increases more then the surface area (where losses are dissipated). Thus, larger cores must be operated at lower power densities.

For the core loss limited case, ΔB_{MAX} may be approximated by assuming a core loss of $100\ mw/cm^3$ – a typical maximum for natural convection cooling. For the core material used, enter the core loss curves at $100\ mw/cm^3$ (Fig. 2-3). Go across to the appropriate switching (ripple) frequency curve, then down to the "Flux Density" scale (actually peak flux density). Double this number to obtain peak-peak flux density, ΔB_{MAX}. If units are in Gauss, divide by 10,000 to convert ΔB_{MAX} to Tesla, then enter this value into the core loss limited Area Product formula.

In filter inductor applications, normally operated in the continuous inductor current mode, the ripple current is usually only 10-20% of the full load dc current. Ferrite cores will usually be limited by saturation flux density, not by core loss, at switching frequencies below 250 kHz. Boost and flyback inductors, and flyback transformers operated in the continuous current mode, where total ripple ampere-turns are a small fraction of full-load ampere-turns, may also be saturation limited. In these situations, it may be possible to reduce size, weight, and/or cost by using core materials that are lossier but have higher saturation flux density, such as Kool-Mu®, or metal alloy laminated cores.

However, when these applications are designed for discontinuous mode operation at full load, ripple ampere-turns are so large that the inductors will almost certainly be core loss limited.

If uncertain whether the application is core loss limited or saturation limited, evaluate both formulae and use the one which results in the largest area product.

These initial estimates of core size are not very accurate, but they do reduce the number of trial solutions that might otherwise be required. The detailed design process provides greater accuracy. In the final analysis, the validity of the design should be checked with a prototype operated in the circuit and in the environment of the application, with the hot spot temperature rise measured by means of a thermocouple cemented alongside the middle of the centerpost.

Inductance calculation

Several methods are in common use for calculating inductance:

Discrete gap length, ℓ_g : The magnetic path length of any core with a discrete gap consists of very high permeability magnetic core material (μ_r = 3000 - -100,000) in series with a small non-magnetic gap (μ_r = 1). In practice, the reluctance of the magnetic material is so small compared to the gap reluctance, that it can usually be neglected. The corrected gap dimensions alone determine the inductance:

$$L = \mu_0 N^2 A_g / \ell_g \times 10^{-2} \text{ Henrys} \qquad \textbf{(3a)}$$

(SI units, dimensions in cm)
A_g = *corrected* gap area (page 5-4)

Effective permeability, μ_e : Whether the gap is discrete or distributed, it is a small total length of non-magnetic material in series with a much greater length of high permeability magnetic material. The actual core can be considered equivalent to a solid homogeneous core with the same overall core dimensions, made entirely of an imaginary material with permeability μ_e, which typically ranges from 10 (for a large gap) to 300 (for a small gap). This concept is most useful for distributed gap powdered metal cores, where the total gap cannot be physically measured.

$$L = \mu_0 \mu_e N^2 A_e / \ell_e \times 10^{-2} \text{ Henrys} \qquad \textbf{(3b)}$$

(SI units, dimensions in cm)

Inductance Factor, A_L, expressed in milliHenrys/1000 turns^2, or nanoHenrys/turn^2, is often stated by the manufacturer for pre-gapped ferrite cores or for distributed-gap powdered metal cores. It provides a convenient method for calculating inductance for an existing gapped core with a given number of turns, but it is awkward for determining the optimum gap length or the optimum effective permeability for best core utilization.

$$L = N^2 A_L \text{ nanoHenrys} \qquad \textbf{(3c)}$$

In the inductor design process, the desired inductance is presumed to be a known circuit value. The optimum gap length, ℓ_g, or effective permeability, μ_e to achieve that inductance is calculated by inverting the preceding formulae.

Design Strategy

The general design procedure that will be followed is:

1. From the circuit design, **define the circuit parameters** including inductance value L, full load dc inductor current I_{FL}, worst case ripple ΔI_{pp}, max peak instantaneous short-circuit current limit I_{SCpk}. absolute loss limit and max temperature rise. *Worst case ripple is at max V_{IN} for buck-derived, at min V_{IN} for boost. Full load inductor current equals load current for buck only. Refer to Section 2.*

2. **Select the core material**. Refer to Section 2.

3. **Determine the maximum flux density and max. flux swing** at which the core will be operated (limited either by saturation or by core loss). Define a conservative saturation limit, B_{MAX} (perhaps 3000Gauss (0.3Tesla) for power ferrite). If the core is saturation limited, B_{MAX} will be reached at I_{SCpk}. With a discrete gapped core, the gap has the most significant influence on the B-H characteristic, linearizing it until well into saturation. Therefore, if the core is saturation limited, maximum flux *swing* ΔB_{MAX} will be in the same proportion to ΔI_{pp} as B_{MAX} is to I_{SCpk}:

$$\Delta B_{MAX} = B_{MAX} \frac{\Delta I_{pp}}{I_{SCpk}} \qquad \textbf{(4)}$$

Divide the calculated ΔB_{MAX} value by two to convert peak-peak flux density to peak, and enter the core loss curve (Fig. 2-3) on the "flux density" axis (really *peak* flux density). At the ripple frequency curve, find the resulting core loss. If the core loss is much less than 100 mw/cm^3, this confirms that the core is probably saturation limited, and the calculated ΔB_{MAX} value is probably valid. But if the indicated core loss is much greater than 100 mw/cm^3, the core will probably be loss limited. ΔB_{MAX} must then be reduced to achieve an acceptable core loss (Step 5). If the core is loss limited, peak flux density at I_{SCpk} will then be less than B_{MAX}.

The approach taken above equates flux density values with currents, based on the presumption of a linear core characteristic. This presumption is quite valid for a gapped ferrite or laminated metal core. On the other hand, powdered metal cores are quite non-linear over a substantial portion of their range. But in high frequency switching power supply applications, these cores will usually be limited by core loss to well below saturation flux density, where linearity is much better. Nevertheless, the determination of core loss and max. permissible flux swing are best accomplished by methods defined by the manufacturers of these cores. (Also, be aware that the permeability quoted by the manufacturer may not apply at the conditions of the application.)

4. Tentatively **select the core shape and size.** Inexperienced designers should use the area product formulae (Eq. 2a, 2b), or manufacturer's guidance. Record the important core dimensions.

5. **Determine loss limit.** First, define the thermal resistance from the data sheet or calculate R_T according to page 4-2. Divide the temperature rise limit by the thermal resistance to calculate the temperature rise loss limit. Compare the temperature rise loss limit with the absolute loss limit, and use whichever is smaller.
   If the core is limited by loss rather than by saturation, initially apportion half of the loss limit to the core and half to the windings. Then apply the core loss limit to the core loss curves to find the ΔB_{MAX} value that will produce that core loss.

6. **Calculate the number of turns,** N, that will provide the desired inductance value when operated at the max flux density swing as defined in Step 3 or Step 5.

$$E = N\frac{d\phi}{dt} = NA_e\frac{dB}{dt} \quad \text{Faraday's Law}$$

$$E = L\frac{\Delta I}{\Delta t}$$

Combining the above (L in μH, A_e in cm):

$$N = \frac{L\Delta I_{MAX}}{\Delta B_{MAX} A_e} \times 10^{-2} \qquad \textbf{(5)}$$

N must then be rounded to an integer value. If N is rounded down to a smaller integer, the core may saturate, or, if the core is loss limited, core loss will be greater than planned. However, winding loss will be reduced. If N is rounded up to a larger integer value, core loss will be reduced, but winding loss increased. When N is a small number of turns, there is a very large increase in winding loss when rounding up vs. rounding down. It may be advantageous to round down to the smaller integral N value if the reduced winding loss outweighs the increased core loss.

When an inductor has multiple windings, the lowest voltage winding with the fewest turns usually dominates the rounding decision. De-optimization caused by rounding sometimes forces the need for a larger core. It may be desirable to alter the turns ratio, or use a smaller inductance value (resulting in greater ripple current) to avoid increased losses or the need for a larger core.

Equation 5 can be applied to any winding provided N, L, and ΔI are all referred to that winding.

After the integer value of N has been established, recalculate ΔB, inverting Eq.5, and determine the resulting core loss.

7. **Calculate the gap length** required to achieve the required inductance, using the N value established in Step 6 (inverting the inductance formula: Eq. 3a, 3b).
   For a discrete gap, the *effective* magnetic path length is the gap, ℓ_g. The center-pole area, A_e must be corrected for the fringing field (Eq. 1a or 1b) to obtain the effective gap area A_g.

$$\ell_g = \mu_0 N^2 \frac{A_g}{L} \times 10^4 \qquad \textbf{(6a)}$$

$$\ell_g = \mu_0 N^2 \frac{A_e}{L}\left(1+\frac{\ell_g}{D_{CP}}\right)^2 \times 10^4 \qquad \text{(6b)}$$

(SI units, L in μH, dimensions in cm)

The solution is a bit messy: Assume a value for ℓ_g in the gap area correction factor term on the right, and calculate a new ℓ_g value. Using this new ℓ_g value for the gap correction, recalculate. Iterate 2-3 times.

For a distributed gap in a powdered metal core, calculate the effective permeability of the composite material required to obtain the desired inductance (or calculate the inductance factor, A_L, discussed earlier):

$$\mu_e = \frac{L\ell_e}{\mu_0 N^2 A_e} \times 10^{-4} \qquad \text{(7)}$$

(L in μH, dimensions in cm)

8. **Calculate the conductor size and winding resistance.** (Refer to the following cookbook sections for details.)
Winding resistance is calculated using the conductor cross-section area and length.
Resistivity of copper:

$$\rho_{cu} = 1.724[1+.0042(T-20)]\times 10^{-6} \quad \Omega\text{-cm}$$

$$\rho_{cu} = 2.30\times 10^{-6} \quad \Omega\text{-cm at } 100°\text{C}$$

dc winding resistance:

$$R_x = \rho_{cu}\, {}^{l_x}\!/\!{}_{A_x} \quad \Omega \qquad \text{(8)}$$

(Dimensions in cm)

9. **Calculate winding loss, total loss, and temperature rise. If loss or temperature rise is too high or too low,** iterate to a larger or smaller core size.

The cookbook design examples which follow will more fully illustrate this design process.

Cookbook Example: Buck Output Filter Inductor

In Section 4, a forward converter transformer was designed for 5V, 50A output. This filter inductor will be designed as the output filter for this same power supply.

1. Define the power supply parameters pertaining to the inductor design. (Vin for the inductor equals Vin for the transformer divided by the 7.5:1 transformer turns ratio):

VIN Range:	13.33 – 25.33 V
Output 1:	5 V
Full Load Current, I_{FL}:	50 A
Circuit Topology:	Forward Converter
Switching Freq, f_S:	200 kHz
Max Duty Cycle:	.405 (at Min VIN)
Min Duty Cycle:	.213 (at Max VIN)
Max Ripple Current, ΔI_{pp}:	50A x 20% = 10 A
Max peak Current, I_{SCpk}:	65 A
Inductance, L:	2.2 μH
Max Loss (absolute):	2.5 W
Max °C Rise:	40°C
Cooling Method:	Natural Convection

2: Select the core material, using guidance from the manufacturer's data sheet.

Core Material: Ferrite, Magnetics Type P

3. Determine max. flux density and max. flux swing at which the core will be operated. A saturation-limited B_{MAX} of 0.3T (3000 Gauss) will be used. If the core is saturation limited, it will be at the B_{MAX} limit when the peak current is at the short-circuit limit. The max. peak-peak flux density swing corresponding to the max. current ripple will then be:

$$\Delta B_{MAX} = B_{MAX}\frac{\Delta I_{pp}}{I_{SCpk}} = 0.3\frac{10}{65} = .046 \text{ Tesla}$$

Dividing the peak-peak flux density swing by 2, the peak flux swing is .023T (230 Gauss). Entering the core loss curve for type P material (page 2-5) at 230 Gauss, and at the 200kHz ripple frequency, the core loss is approximately 4mw/cm^3. This is so much less than the 100 mw/cm^3 rule of thumb that core loss will be almost negligible, and core operation will be saturation limited at I_{SCpk}. The maximum flux density swing, ΔB_{MAX}, is therefore .046T as previously calculated.

4. Tentatively select core shape and size, using guidance from the manufacturer's data sheet or using the area product formula given previously.

Core type, Family: E-E core – ETD Series

Using the saturation limited Area Product formula, with B_{MAX} = 0.3T and K1 =.03, an Area Product of 0.74 cm^4 is required. An ETD34 core size will be used, with AP = 1.21 cm^4 (with bobbin).

Core Size: 34mm -- ETD34

For the specific core selected, note:

Effective core Area, Volume, Path Length, center-pole diameter (cm):

A_e: 0.97 cm²
V_e: 7.64 cm³
ℓ_e: 7.9 cm
D_{CP}: 1.08 cm

Window Area, Breadth, Height, Mean Length per Turn (with bobbin):

A_w: 1.23 cm²
bw: 2.10 cm
hw: 0.60 cm
MLT: 6.10 cm

5. Define R_T and Loss Limit. Apportion losses to the core and winding. Thermal resistance from the data sheet is 19°C/Watt. Loss limit based on max. temperature rise:

Plim = °Crise/R_T: = 40/19 = 2.1 Watts

Compared to the absolute loss limit of 2.5W (Step 1), the temperature rise limit of 2.1W applies. Core loss is 4 mW/cm³ (Step 3):

PC = mW/cm³ x V_e = 4 x 7.64 = 30mW

Therefore, winding loss can be as much as 2 Watts. However, since the core is larger than the Area Product calculation suggests, it should be possible to reduce the winding loss.

6. Calculate the number of turns that will provide the desired inductance value:

$$N = \frac{L\Delta I_{MAX}}{\Delta B_{MAX} A_e} \times 10^{-2} \qquad (5)$$

(L in μH, dimensions in cm)

$$N = \frac{2.2 \cdot 10}{.046 \cdot 0.97} \times 10^{-2} = 4.93 \rightarrow 5 \text{ Turns}$$

7. Calculate the gap length that will achieve the required inductance value:

$$\ell_g = \mu_0 N^2 \frac{A_e}{L}\left(1 + \frac{\ell_g}{D_{CP}}\right)^2 \times 10^4 \qquad (6b)$$

(L in μH, dimensions in cm)

$$\ell_g = 4\pi \times 10^{-7} \cdot 5^2 \frac{0.97}{2.2}\left(1 + \frac{\ell_g}{1.08}\right)^2 \times 10^4$$

$$\ell_g = 0.192 \text{ cm}$$

8. Calculate the conductor size, winding resistance, losses, and temperature rise.

From Step 4, window breadth, b_w = 2.10cm, and height, h_w = 0.60cm. The winding will consist of 5 turns (5 layers) of copper strip, 2.0cm wide, spiral wound, with .05mm (2 mil) low voltage insulation between layers.

At 50A full load current, 450 A/cm² requires a conductor area of 0.111cm², Dividing this conductor area by the 2.0cm width requires a thickness of .0555cm. Five layers, including .005cm insulation between layers, results in a total winding height of 0.3cm, half the available window height.

To reduce losses, increasing copper thickness to 0.1cm results in a total winding height of .525cm, and a conductor area of 0.2cm². Five turns with a mean length/turn = 6.1cm results in a total winding length of 30.5cm. Winding resistance:

$$R_{dc} = \rho \frac{l}{A} = 2.3 \times 10^{-6} \cdot \frac{30.9}{0.2} = .000355\Omega$$

DC loss: $50^2 \cdot .000355 = 0.89$ Watts

With reference to Section 3-4, the ac loss is calculated. Skin depth D_{PEN} = .017cm at 200kHz. With a conductor thickness of 0.1cm, Q = 0.1/.017 = 5.9. Entering Dowell's curves, page 3-4, with Q=6 and 5 layers, R_{AC}/R_{DC} is approximately 100, so that R_{AC} = .035Ω.

The rms value of the triangular ripple current waveform equals $\Delta I_{pp}/\sqrt{12}$ Since max ΔI_{pp} is 10A, $I_{rms} = 10/\sqrt{12} = 2.9$A.

Therefore the ac loss is:

P_{Lac} = I²R = 2.9²·.035 = 0.29 Watts

Total winding loss dc plus ac is:

P_W = 0.89 + 0.29 = 1.18 Watts

9. Since the core loss is only 30mW, the total loss, 1.21W, is considerably less than the 2.1W limit originally calculated. The windings operate with only 250 A/cm² at full load, accounting for the reduced loss. This is because the ETD34 core has an Area Product 65% greater than the calculated requirement. A smaller core could possibly have been used. However, the ETD34 core provides improved power supply efficiency.

Coupled Filter Inductors

Buck-derived converters with multiple outputs often use a single filter inductor with multiple coupled windings, rather than individual inductors for each output. The design process for a coupled inductor, discussed in Ref. R5 is essentially the same as for a single-winding inductor.

The process is simplified by assuming that all windings are normalized and combined with the lowest voltage winding. Finally, divide the copper cross-section area into the actual multiple windings. Each winding will then occupy an area (A_WN) proportional to its power output.

Boost Inductor

In a simple boost application, the inductor design is essentially the same as for the buck converter discussed previously.

In switching power supplies, boost topologies are widely used in Power Factor Correction applications and in low voltage battery power sources. Otherwise, the boost configuration is rarely used.

In a PFC application, boost inductor design is complicated by the fact that the input voltage is not dc, but the continuously varying full-wave rectified line voltage waveform. Thus, as V_{IN} changes with the line voltage waveform, the high frequency waveforms must also change. High frequency ripple current, flux swing, core loss and winding loss all change radically throughout the rectified line period.

The situation is further complicated by the fact that in different PFC applications, the boost topology may be designed to operate in one of a wide variety of modes:

- *Continuous mode, fixed frequency*
- *Continuous mode, variable frequency*
- *At the mode boundary, variable frequency*
- *Discontinuous mode, fixed frequency*
- *Discontinuous mode, variable frequency*
- *Continuous mode, transitioning to discontinuous during the low current portion of the line cycle, and at light loads.*

As in the buck-derived applications, the limiting factors for the boost inductor design are (a) losses, averaged over the rectified line period, and/or (b) core saturation at maximum peak current.

Worst case for core saturation is at maximum peak current, occurring at low line voltage at the peak of the rectified line voltage waveform. This is usually easy to calculate, regardless of the operating mode.

Note that the simple boost topology has no inherent current limiting capability, other than the series resistance of the line, rectifiers, and bulk filter capacitor. Thus, the boost inductor *will saturate momentarily* during start-up, while the bulk capacitor charges. The resulting inrush current is basically the same as with a simple capacitor-input filter, and is usually acceptable in low power applications. In high power applications, additional means of inrush current limiting is usually provided – a thermistor or an input buck current limiter. While saturation may be permissible during startup, the circuit must be designed so that the inductor does not saturate during worst-case normal operating conditions.

Calculating the losses averaged over the rectified line period is a difficult task. Averaged losses can be approximated by assuming V_{IN} is constant, equal to the rms value of the actual rectified line voltage waveform.

Because input current is greatest at low input voltage, low frequency winding losses are greatest at low V_{IN}.

For discontinuous mode operation, $\Delta I_{p\text{-}p}$ is greatest at low V_{IN}. Therefore, core loss and ac winding loss are worst case at low line.

However, when the boost topology is operated in the continuous mode, max $\Delta I_{p\text{-}p}$ and worst case core loss and ac winding loss occur when V_{IN} equals one-half of V_O. But since $\Delta I_{p\text{-}p}$ is usually very much smaller than low frequency current, core loss and ac winding loss are usually negligible in continuous mode operation.

Because the boost inductor design follows the same pattern as the output filter inductor design previously covered, a cookbook example of boost inductor design is not given. It is left to the designer to determine the worst-case current values governing the design.

Flyback Transformer Design

Figures 5-6 and 5-7 show the inductor current waveforms for continuous mode and discontinuous mode operation. All currents are normalized to their ampere-turn equivalent by multiplying primary and secondary currents by their respective winding turns. The ampere-turns driving the core are thus proportional to the normalized currents shown.

The design of the flyback transformer and calculation of losses requires definition of duty cycle, D, from which the transformer turns ratio, n, is calculated according to the relationship:

$$n = \frac{V_{IN}}{V_O'}\frac{D}{1-D} \quad ; \quad D = \frac{nV_O'}{V_{IN}+nV_O'} \qquad (9)$$

where V_O' equals output voltage plus rectifier, switch and IR drops referred to the secondary. The above relationship applies to discontinuous mode operation, and for discontinuous mode operation only at the mode boundary

Theoretically, a transformer-coupled flyback circuit can function with any turns ratio, regardless of V_{IN} or V_O. However, it functions best, avoiding high peak currents and voltages, when n is such that D is approximately 0.5. (for discontinuous mode operation, at the mode boundary.) Circuit considerations and device ratings may dictate a turns ratio that results in a duty cycle other than 0.5. The turns ratio determines the trade-off between primary and secondary peak voltages and peak currents. For example, reducing n reduces the duty cycle, reduces peak switch voltage and peak rectifier current, but increases peak switch current and peak rectifier voltage.

Waveform Definitions: Before flyback transformer design can be completed, the dc, ac and total rms current components of each waveform must be calculated. Current values must be calculated at each of the differing worst-case conditions relevant to core saturation, core loss and winding loss.

D_P Duty Cycle, primary (switch) waveform
D_S Duty Cycle, secondary (diode) waveform
$D_S = (1\text{-}D_P)$ In the continuous mode and at the discontinuous mode boundary.
I_{pk} Ripple current max peak
I_{min} Ripple current min peak
ΔI_{pp} pk-pk Ripple current, $I_{pk} - I_{min}$
I_{pa} average value of trapezoidal peak: $I_{pa} = (I_{pk} + I_{min})/2$.

The following equations can be used to calculate the dc, rms, and ac values of trapezoidal waveforms (continuous mode operation – Fig. 5-6). They also apply to triangular waveforms (discontinuous mode – Fig. 5-8), by setting I_{min} to zero:

$$I_{dc} = D\frac{\left(I_{pk}+I_{\min}\right)}{2} = D\cdot I_{pa} \qquad (10)$$

$$I_{rms} = \sqrt{D\left[\left(I_{pk}\times I_{\min}\right)+\frac{1}{3}\left(I_{pk}-I_{\min}\right)^2\right]} \qquad (11)$$

For trapezoidal waveforms, Equation 11 can be simplified by ignoring the slope of the waveform top. If $\Delta I_{pp} = 0.5\, I_{pa}$, the error is only 1%. If $\Delta I_{pp} = I_{pa}$, the error is 4%.

$$I_{rms} \approx \sqrt{D\cdot I_{pa}^{\ 2}} \qquad (11a)$$

For triangular waveforms, Eq. 11 becomes:

$$I_{rms} = \sqrt{\frac{D}{3}I_{pk}^{\ 2}} \qquad (11b)$$

For all waveforms:

$$I_{ac} = \sqrt{I_{rms}^{\ 2} - I_{dc}^{\ 2}} \qquad (12)$$

Switching transition times are governed by transformer leakage inductance as well as by transistor and rectifier switching speeds. Thus it is important to minimize leakage inductance by using cores with long, narrow windows, and by interleaving the windings. Although switching transitions result in switch and rectifier losses, they have little effect on transformer loss.

Continuous Mode Operation

With continuous mode operation, core loss is usually not significant because the ac ripple component of the *total* inductor ampere-turns is small compared with the full load dc component. But currents in the individual windings switch on and off, transferring ampere-turns back and forth from primary to secondary(s), as shown in Fig. 5-6. This results in very large ac current components in the windings that will likely result in significant high frequency winding losses.

The secondary current dc component is equal to output current, regardless of V_{IN}. At low V_{IN} the primary dc and peak currents and the total inductor ampere-turns are greatest. Thus, the worst-case condition for core saturation and winding losses occurs at low V_{IN}.

On the other hand, the ac ripple component of the total inductor ampere-turns, and thus core loss, is greatest at high V_{IN}. But since core loss is usually negligible with continuous mode operation, this has little significance.

Cookbook Example (Continuous Mode):

1. Define the power supply parameters pertaining to the flyback transformer design.

VIN:	28 ± 4 V
Output 1:	5 V
Full Load Current, IFL:	10 A
Circuit Topology:	Flyback, Continuous Mode
Switching Freq, fs:	100 kHz
Desired Duty Cycle:	0.5 at 28V input
Max Ripple Current, ΔI_{pp}:	5 A @ 32V (secondary)
Peak Short-circuit Current:	25A (secondary)
Secondary Inductance, L:	6.8 μH (D=0.5, ΔI_{pp} = 5A)
Max Loss (absolute):	2.0 W
Max Temperature Rise:	40°C
Cooling Method:	Natural Convection

Preliminary Calculations: The turns ratio can be defined at nominal V_{IN} = 28V and the desired duty cycle of 0.5:

$$n = \frac{V_{IN}}{V_O'} \frac{D}{1-D} = \frac{28}{5+0.6} \frac{0.5}{1-0.5} = 5$$

Before calculating winding losses, worst-case dc and ac current components, occurring at low V_{IN}, must be defined. First, the duty cycle, D_P is defined at low V_{IN}:

$$D_{P24} = \frac{nV_O'}{V_{IN} + nV_O'} = \frac{5(5+0.6)}{24+5(5+0.6)} = 0.538$$

$$D_{S24} = 1 - D_{P24} = 0.462$$

Because the duty cycle and the turns ratio could possibly be changed to optimize the windings, current calculations are deferred until later.

2: Select the core material, using guidance from the manufacturer's data sheet.

Core Material: Ferrite, Magnetics Type P

3. Determine max. flux density and max. flux swing for core operation. A saturation-limited B_{MAX} of 0.3T (3000 Gauss) will be used. If the core is saturation limited, B will reach B_{MAX} when peak current reaches the short-circuit limit. Assuming reasonable linearity of the gapped core B-H characteristic, ΔB_{MAX} with max. current ripple (at 32V) will be:

$$\Delta B_{MAX} = B_{MAX} \frac{\Delta I_{pp}}{I_{SCpk}} = 0.3 \frac{5}{25} = .06 \text{ Tesla} \quad (4)$$

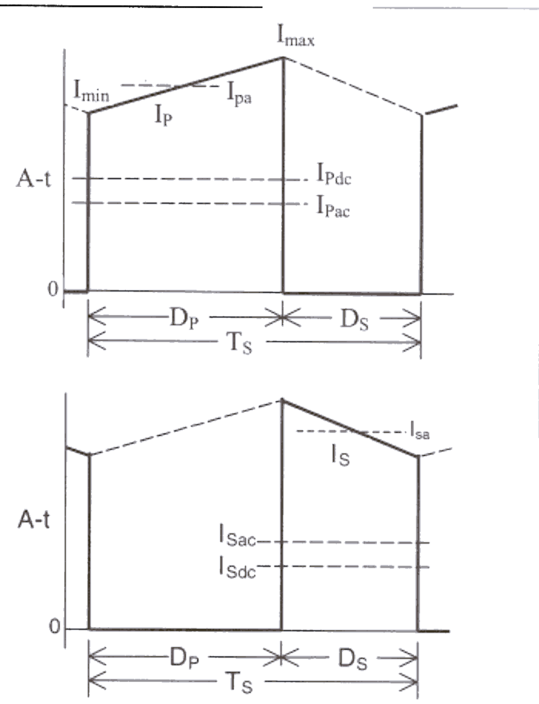

Figure 5-6 Flyback Waveforms, Continuous

Dividing the peak-peak flux density swing by 2, the peak flux swing is .03T (300 Gauss). Entering the core loss curve for type P material (page 2-5) at 300 Gauss, and at 100kHz ripple frequency, the core loss is approximately 2.6 mw/cm^3. This is so much less than the 100 mw/cm^3 rule of thumb that core loss is negligible. Thus, B_{MAX} is saturation limited at I_{SCpk} of 25A, and ΔB_{MAX} , is limited to only .06T, corresponding to $\Delta I_{p\text{-}p}$ of 5Amp.

4. Tentatively select the core shape and size, using guidance from the manufacturer's data sheet or using the area product formula.

Core type, Family: E-E core – ETD Series

Using the saturation limited Area Product formula, with B_{MAX} = 0.3T and K1 =.0085, an Area Product of 1.08 cm^4 is required. An ETD34 core size will be used, with AP = 1.21 cm^4 (with bobbin).

Core Size: 34mm -- ETD34

For the specific core selected, note:

Effective core Area, Volume, Path Length, Center-pole diameter (cm):

A_e: 0.97 cm^2
V_e: 7.64 cm^3
ℓ_e: 7.9 cm
D_{CP}: 1.08 cm

Window Area, Breadth, Height, Mean Length per Turn (with bobbin):

A_w: 1.23 cm^2
bw: 2.10 cm
hw: 0.60 cm
MLT: 6.10 cm

5. Define R_T and Loss Limit. Apportion losses to the core and winding. Thermal resistance from the core data sheet is 19°C/Watt. Loss limit based on max. temperature rise:

Plim = °Crise/R_T: = 40/19 = 2.1 Watts

Since this exceeds the 2.0W absolute loss limit from Step 1, The 2.0W limit applies.

Core loss is:

PC = mW/cm^3 x V_e = 2.6 x 7.64 = 20mW

Therefore, core loss is negligible. The entire 2.0 Watt loss limit can be allocated to the winding.

6. Calculate the number of secondary turns that will provide the desired inductance value:

$$N = \frac{L\Delta I_{MAX}}{\Delta B_{MAX} A_e} \times 10^{-2} \qquad (5)$$

(L in µH, dimensions in cm)

$$N_S = \frac{6.8 \cdot 5}{.06 \cdot 0.97} \times 10^{-2} = 5.84 \rightarrow 6 \text{ Turns}$$

$$N_P = N_S \times n = 6 \times 5 = 30 \text{ Turns}$$

7. Calculate the gap length to achieve the inductance value with minimum N.

$$\ell_g = \mu_0 N^2 \frac{A_e}{L}\left(1 + \frac{\ell_g}{D_{CP}}\right)^2 \times 10^4 \qquad (6b)$$

(L in µH, dimensions in cm)

$$\ell_g = 4\pi \times 10^{-7} \cdot 5^2 \frac{0.97}{6.8}\left(1 + \frac{\ell_g}{1.08}\right)^2 \times 10^4$$

$$\ell_g = .080 \text{ cm}$$

8. Calculate the conductor sizes and winding resistances:

From Step 4, window breadth, b_w = 2.10cm, and height, h_w = 0.60cm. A creepage allowance of 0.3 cm is necessary at each end of the windings. Winding width is 2.10cm minus (2x0.3) = 1.5cm.

Secondary Side – V_{IN} = 24V, D_S = 0.462 (Eq. 10, 11a, 12)

Output dc Current, I_{Sdc}: **10** A

Avg peak Current, I_{Spa}: $\frac{I_{DC}}{D_S} = \frac{10}{.462} = 21.65$ A

rms Current, I_{Srms}: $\sqrt{D_S \cdot I_{pa}^2} = 14.7$ A

ac Current, I_{Sac}: $\sqrt{I_{rms}^2 - I_{dc}^2} = 10.77$ A

The secondary consists of 6 turns (6 layers) of copper strip, 1.5cm wide and .015 cm thick, spiral wound. Conductor area is .015x1.5 = .0225 cm^2. Current density is 14.7A/.0225 = 650 a/cm^2.

Six layers, including .005cm (2 mil) low voltage insulation between layers, results in a total winding height of 0.12cm.

Six turns with mean length/turn = 6.1cm results in a total winding length of 36.6cm. Winding resistance:

$$R_{dc} = \rho\frac{l}{A} = 2.3 \times 10^{-6} \cdot \frac{36.6}{.0225} = .0037\Omega$$

Calculating ac resistance: D_{PEN} at 100kHz = .024cm. With a conductor thickness of .015cm, Q =.015/.024 = 0.625. Entering Dowell's curves, page 3-4, with Q = 0.625 and 6 layers, R_{AC}/R_{DC} is approximately 1.6.

$$R_{ac} = R_{dc} \times 1.6 = .0037\Omega \times 1.6 = .0059\Omega$$

Primary Side – V_{IN} = 24V, D_P = 0.538 (Eq. 10, 11a, 12)

Note that the primary and secondary *average peak ampere-turns* are always equal, and together constitute the dc ampere-turns driving the inductor core. Thus the primary avg. peak current, $I_{Ppa} = I_{Spa}/n$.

avg. peak Current, I_{Ppa}: $I_{Spa}/n = 21.65/5 = 4.33$ A

dc Current, I_{Pdc}: $D \cdot I_{Ppa} = 0.538 \times 4.33 = 2.33$A

rms Current, I_{Prms}: $\sqrt{D_P \cdot I_{Ppa}^2} = 3.18$ A

ac Current, I_{Pac}: $\sqrt{I_{rms}^2 - I_{dc}^2} = 2.16$A

peak SC Current, I_{SCpk}: 25/n = 5 A

The primary winding consists of 30 turns of Litz wire with OD of 0.127cm, in three layers, 10 turns in each layer. Litz wire OD enables 10 turns to fit across the 1.5cm winding breadth. The height of the 3-layer primary is 3x0.127 = 0.381cm.

The Litz wire consists of 150 strands of #40AWG wire (**OD=.0081cm**). From the wire tables, #40AWG wire has a resistance of .046 Ω/cm divided by 150 strands, resulting in a resistance of **.00031 Ω/cm** at 100°C.

The wire length equals 30 turns times MLT of 6.1cm = **183cm**.

Primary winding resistance:

$$\boldsymbol{R_{dc} = .00031\,\Omega / \mathrm{cm} \times 183\mathrm{cm} = .0567\,\Omega}$$

To calculate the ac resistance, the 150 strand #40AWG Litz wire is approximately equivalent to a square array 12 wide by 12 deep (square root of 150 wires). There are therefore a total of **36 layers** of #40AWG wire (3 layers times 12).

Center-to-center spacing of the #40AWG wires equals the winding width of 1.5cm divided by 120 (10 Litz wires times 12 wide #40AWG wires within the Litz), a spacing, *s*, of **.0125cm.**

From Reference R2, pg 9, the effective layer thickness equals:

$$\boldsymbol{0.83d(d/s)^{1/2} = 0.83 \cdot .0081\sqrt{.0081/.0125}}$$

$$\boldsymbol{= .0054\mathrm{cm}}$$

Therefore, referring to Figure 3-5,

$$\boldsymbol{Q = .0054cm / D_{PEN} = .0054/.024 = .225}$$

and with 36 layers, $\boldsymbol{R_{AC}/R_{DC}}$ **= 1.6**, and

$$\boldsymbol{R_{ac} = R_{dc} \times 1.6 = .0567 \times 1.6 = .090\Omega}$$

9. Calculate winding loss, total loss, and temperature rise:

Secondary dc loss:

$$\boldsymbol{P_{Sdc} = I_{dc}^2 R_{dc} = 10^2 \cdot .0037 = 0.37} \text{ Watts}$$

Secondary ac loss:

$$\boldsymbol{P_{Sac} = I_{Sac}^2 \cdot R_{ac} = 10.77^2 \cdot .0059 = 0.68} \text{W}$$

Total secondary winding loss -- dc plus ac is:

$\boldsymbol{P_{Sw} = 0.37 + 0.68 = 1.05}$ Watts

The available winding height could permit a thicker secondary conductor. This would reduce dc loss, but the resulting increase in ac loss because of the larger Q value would exceed the dc loss reduction.

Primary dc loss (R_{dc} = .0567Ω):

$$\boldsymbol{P_{Pdc} = I_{Pdc}^2 R_{dc} = 2^2 .0567 = 0.225} \text{ Watts}$$

Primary ac loss (R_{ac} = .090Ω):

$$\boldsymbol{P_{Pac} = I_{Pac}^2 R_{ac} = 2.16^2 \times .090 = 0.42} \text{ W}$$

Total primary winding loss -- dc plus ac is:

$\boldsymbol{P_{Pw} = 0.225 + 0.42 = 0.645}$ Watts

Total winding loss:

$\boldsymbol{P_W = 1.05 + 0.645 = 1.695}$ Watts

Since the core loss is only 20mW, the total loss, 1.71W, is within than the 2.0W absolute loss limit.

The total winding height, including .02cm isolation: 0.12 + 0.381 + .02 = **0.521cm**, within the available winding height of 0.60cm.

Mutual inductance of 6.8µH seen on the secondary winding translates into 170µH referred to the primary ($L_P = n^2 L_S$).

Leakage inductance between primary and secondary calculated according to the procedure presented in Reference R3 is approximately 5µH, referred to the primary side. Interwinding capacitance is approximately 50pF.

If the windings were configured as an interleaved structure (similar to Figure 4-1), leakage inductance will be more than halved, but interwinding capacitance will double. The interleaved structure divides the winding into two sections, with only half as many layers in each section. This will reduce R_{ac}/R_{dc} to nearly 1.0 in both primary and secondary, reducing ac losses by 0.35W, and reducing the total power loss from 1.71W to 1.36W. The secondary copper thickness could be increased, further reducing the losses.

Discontinuous Mode Operation

Discontinuous mode waveforms are illustrated in Figure 5-7. By definition, the total ampere-turns dwell at zero during a portion of each switching period. Thus, in the discontinuous mode there are three distinct time periods, t_T, t_R, and t_0, during each switching period. As the load is increased, peak currents, t_T, and t_R increase, but t_0 decreases. When t_0

becomes zero, the mode boundary is reached. Further increase in load results in crossing into continuous mode operation. This is undesirable because the control loop characteristic suddenly changes, causing the control loop to become unstable.

In the discontinuous mode, *all* of the energy stored in the inductor ($\frac{1}{2}LI_{pk}^2$) is delivered to the output during each cycle. This energy times switching frequency equals output power. Therefore, if frequency, L, and V_O are held constant, $\frac{1}{2}LI_{pk}^2$ does not vary with V_{IN}, but is proportional to load current only, and I_{pk} is proportional to the square root of load current. However, V_{IN}, n, D and L collectively *do* determine the *maximum* stored energy and therefore the maximum power output at the mode boundary.

The circuit should be designed so that the peak short-circuit current limit is reached just before the mode boundary is reached, with turns ratio, duty cycle and inductance value designed to provide the necessary full load power output at a peak current less than the current limit.

The circuit design can never be completely separated from the design of the magnetic components. This is especially true at high frequencies where the small number of secondary turns can require difficult choices. The ideal design for a discontinuous mode flyback transformer might call for a secondary with 1½ turns. The choice of a 1 turn or 2 turn secondary may result in a size and cost increase, unless the turns ratio and duty cycle are changed, for example.

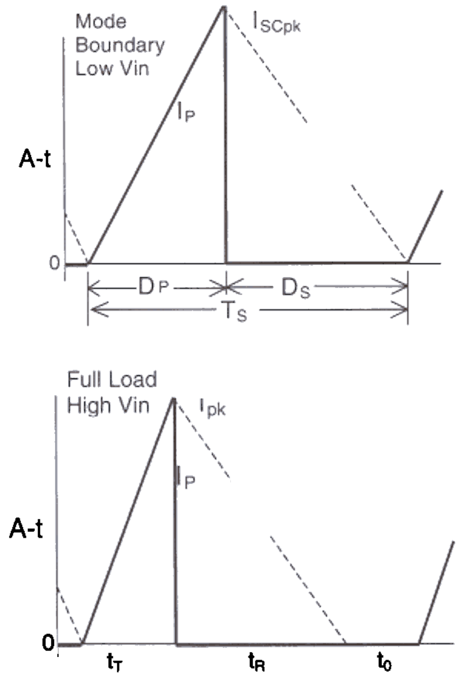

Figure 5-7 Discontinuous Waveforms

Cookbook Example (Discontinuous Mode):

1. Define the power supply parameters pertaining to the flyback transformer design.

VIN:	28 ± 4 V
Output 1:	5 V
Full Load Current, IFL:	10 A
Short Circuit Current:	12 A
Circuit Topology:	Flyback, Discontinuous
Switching Freq, fs:	100 kHz
Desired Duty Cycle:	0.5 at 24V, mode boundary
Estimated I_{SCpk}:	45 A (secondary)
Est. Sec. Inductance:	0.62µH (D=0.5, ΔI_{p-p}=45A)
Max Loss (absolute):	2.0 W
Max Temperature Rise:	40°C
Cooling Method:	Natural Convection

Preliminary Calculations: The turns ratio is defined based on min V_{IN} (24V) and V_O' (5.6V) and the desired duty cycle of 0.5 at the mode boundary:

$$n = \frac{V_{IN}}{V_O'}\frac{D}{1-D} = \frac{24}{5+0.6}\frac{0.5}{1-0.5} = 4.28 \to 4$$

Turns ratio, n, is rounded down to 4:1 rather than up to 5:1 because: (a) 4:1 is closer, (b) peak output current is less, reducing the burden on the output filter capacitor, and (c) primary switch peak voltage is less. The duty cycle at the mode boundary is no longer 0.5, and must be recalculated:

$$D_{P24} = \frac{V_O' \cdot n}{V_{IN} + V_O' \cdot n} = \frac{5.6 \times 4}{24 + 5.6 \times 4} = 0.483$$

$$D_{S24} = 1 - D_{P24} = 0.517$$

The peak secondary current at the mode boundary is:

$$I_{SCdc} = I_{SCpk}\frac{D_{S24}}{2}$$

$$I_{SCpk} = I_{SCdc}\frac{2}{D_{S24}} = 12\frac{2}{0.517} = 46.4 \text{ A}$$

The peak short-circuit current limit on the primary side should therefore be set slightly below 11.6A (=46.4/n).

The inductance value required for the secondary current to ramp from 46.4A to zero at the mode boundary is:

$$L = V_O'\frac{\Delta t}{\Delta i} = V_O'\frac{T \cdot D_{S24}}{\Delta i}$$

$$L = 5.6\frac{10 \cdot 0.517}{46.4} = 0.624 \quad \mu H$$

Before calculating winding losses, it is necessary to define worst-case dc and ac current components, occurring at low V_{IN}. Because the turns ratio and the duty cycle could possibly be changed to optimize the windings, current calculations will be deferred until later.

2: Select the core material, using guidance from the manufacturer's data sheet.

Core Material: Ferrite, Magnetics Type P

3. Determine max. flux density and max. flux swing at which the core will be operated. A saturation-limited B_{MAX} of 0.3T (3000 Gauss) will be used. In the discontinuous mode, the current is zero during a portion of each switching period, by definition. Therefore ΔI always equals I_{peak} and since they are proportional, ΔB must equal B_{peak}. ΔB_{MAX} and B_{MAX} occur at low V_{IN} when the current is peak short-circuit limited. To determine if ΔB_{MAX} is core loss limited, enter the core loss curve for type P material at the nominal 100mw/cm^3 loss limit, and at 100kHz ripple frequency, the corresponding max. peak flux density is 1100 Gauss. Multiply by 2 to obtain peak-peak flux density swing ΔB_{MAX} of 2200 Gauss, or 0.22 Tesla. Since in the discontinuous mode, B_{MAX} equals ΔB_{MAX}, then B_{MAX} is also limited to 0.22T, well short of saturation. Thus, in this application, the core is loss-limited at ΔB_{MAX} = 0.22T, corresponding to $\Delta I_{p-p} = I_{SCpk}$ = 46A.

4. Tentatively select the core shape and size, using guidance from the manufacturer's data sheet or using the area product formula.

Core type, Family: E-E core – ETD Series

Using the loss limited Area Product formula, with ΔB_{MAX} = 0.22T and K2 =.006, an Area Product of 0.31 cm^4 is required. An ETD24 core size will be used, with AP = 0.37 cm^4 (with bobbin).

Core Size: 24mm – ETD24

For the specific core selected, note:

Effective core Area, Volume, Path Length, Center-pole diameter (cm):

A_e: 0.56 cm^2
V_e: 3.48 cm^3
ℓ_e: 6.19 cm
D_{CP}: 0..85 cm

Window Area, Breadth, Height, Mean Length per Turn (' indicates reduced dimensions with bobbin):

Aw / Aw': 1.02 / 0.45 cm^2
bw / bw': 2.07 / 1.72 cm
hw / hw': 0.50 / 0.38 cm
MLT: 4.63 cm

5. Define R_T and Loss Limit, and apportion losses to the core and winding. Thermal resistance from the core data sheet is 28°C/Watt. Loss limit based on max. temperature rise:

Plim = °Crise/R_T: = 40/28 = 1.42 Watts

Since this is less than the 2.0W absolute loss limit from Step 1, The 1.42W limit applies.

Preliminary core loss calculation:

Pc = mW/cm^3 x V_e = 100 x 3.48 = 350 mW

6. Calculate the number of secondary turns that will provide the desired inductance value:

$$N = \frac{L\Delta I_{MAX}}{\Delta B_{MAX} A_e} \times 10^{-2} \qquad (5)$$

(L in μH, dimensions in cm)

$$N_S = \frac{0.63 \cdot 46}{.22 \cdot 0.56} \times 10^{-2} = 2.35 \rightarrow 2 \text{ Turns}$$

$$N_P = N_S \times n = 4 \times 2 = 8 \text{ Turns}$$

Because N_S was rounded down from 2.35 to 2 turns, flux density swing is proportionately greater than originally assumed:

$$\Delta B_{MAX} = 0.22\frac{2.35}{2} = 0.258 \quad \text{Tesla}$$

Divide by 2 to obtain peak flux density swing and enter the core loss curve at 0.13T (1300 Gauss) to obtain a corrected core loss of 160mW/cm^3. Multiply by Ve = 3.48 cm for a corrected core loss of 560 mW.

If N_S had been rounded up to 3 turns, instead of down to 2 turns, core loss would be considerably less, but winding loss would increase by an even greater amount, and the windings might not fit into the available window area..

7. Calculate the gap length to achieve the inductance value:

$$\ell_g = \mu_0 N^2 \frac{A_e}{L}\left(1+\frac{\ell_g}{D_{CP}}\right)^2 \times 10^4 \qquad (6b)$$

(L in μH, dimensions in cm)

$$\ell_g = 4\pi\times10^{-7}\cdot 2^2\,\frac{0.56}{0.62}\left(1+\frac{\ell_g}{0.95}\right)^2 \times 10^4$$

$$\ell_g = .050 \text{ cm}$$

8. Calculate the conductor sizes and winding resistances:

From Step 4, window breadth, b_w = 1.72 cm, and height, h_w = 0.38cm. A creepage allowance of 0.3 cm is necessary at each end of the windings. Winding width is 1.72 cm minus (2x0.3) = 1.12cm.

Secondary side: V_{IN} = 24V, D_S = 0.517 (Eq. 11b, 12)

Output dc Current, I_{Sdc}: **12** A (Short Circuit)

Peak S.C. Current, I_{Spk}: **46.4** A

rms Current, I_{Srms}: $\sqrt{\frac{D_S}{3}\cdot I_{pk}^{\;2}}$

$$= \sqrt{\frac{0.517}{3}46.4^2} = 19.2 \text{ A}$$

ac Current, I_{Sac}: $\sqrt{I_{rms}^{\;2} - I_{dc}^{\;2}} = 15$ A

Secondary conductor area for 450 A/cm² requires 19.2A/450 = .043 cm² (AWG 11). This is implemented with copper strip 1.12cm wide and .038 cm thickness, 2 turns, spiral wound.

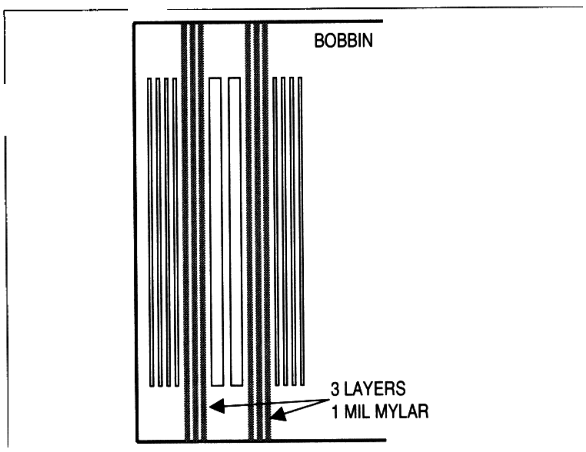

Figure 5-8 – Interleaved Flyback Windings

Two turns -- two layers, including .005cm (2 mil) low voltage insulation between layers, results in a total winding height of .081cm.

Two turns with mean length/turn = 4.6 cm results in a total winding length of 9.2 cm.

Secondary dc resistance:

$$R_{dc} = \rho\frac{l}{A} = 2.3\times10^{-6}\cdot\frac{9.2}{.043} = .00049\Omega$$

Calculating ac resistance: D_{PEN} at 100kHz = .024cm. With a conductor thickness of .033cm, Q =.038/.024 = 1.6. Entering Dowell's curves, page 3-4, with Q = 1.6 and 2 layers, R_{AC}/R_{DC} is approximately 2.5.

Secondary ac resistance (non-interleaved):

$$R_{ac} = R_{dc}\times 2.0 = .00049\Omega\times 2.5 = .00122\Omega$$

If the windings are interleaved, forming two winding sections as shown in Fig. 5-8, there is one secondary turn in each section. Entering Dowell's curves with Q = 1.6 and one layer, R_{AC}/R_{DC} is 1.5.

Secondary ac resistance (interleaved):

$$R_{ac} = R_{dc}\times 1.5 = .00049\Omega\times 1.5 = .0007\Omega$$

Primary side: V_{IN} = 24V, D_P = 0.483 (Eq. 11b, 12)

Note that primary and secondary *peak ampere-turns* are always equal. Thus the primary peak current, $I_{Ppk} = I_{Spk}/n$.

peak S.C.Current, I_{Ppk} $\;I_{Spk}/n = 46.4/4 = 11.6$ A

dc Current, I_{Pdc}: $I_{Ppk}\frac{D_P}{2} = 11.6\frac{0.483}{2} = 2.8$ A

rms Current, I_{Prms}: $\sqrt{\frac{D_P}{3}I_{Ppk}^{\;2}}$

$$= \sqrt{\frac{0.483}{3}11.6^2} = 4.65 \text{ A}$$

ac Current, I_{Pac}: $\sqrt{I_{rms}^{\;2} - I_{dc}^{\;2}} = 3.71$ A

Primary conductor area for 450 A/cm² requires 4.65A/450 = .010 cm² (AWG 17). This is implemented with copper strip 1.12 cm wide and .009 cm thickness, 8 turns, spiral wound.

Eight layers, including .005cm (2 mil) low voltage insulation between layers results in a total winding height of 8x.014 = 0.112cm.

Eight turns with mean length/turn = 4.6 cm results in a total winding length of 36.8 cm.

Primary dc resistance:

$$R_{dc} = \rho \frac{l}{A} = 2.3 \times 10^{-6} \cdot \frac{36.8}{.01} = .0085\Omega$$

Calculating ac resistance: D_{PEN} at 100kHz = .024cm. With a conductor thickness of .009cm, Q =.009/.024 = .375. Entering Dowell's curves, page 3-4, with Q = .375 and 8 layers, R_{AC}/R_{DC} is approximately 1.2.

Primary ac resistance (non-interleaved):

$$R_{ac} = R_{dc} \times 1.2 = .0085\Omega \times 1.2 = .01\Omega$$

With the interleaved structure, there are only 4 layers in each winding section. Entering Dowell's curves, page 3-4, with Q = .375 and 4 layers, R_{AC}/R_{DC} is 1.0.

Primary ac resistance (interleaved):

$$R_{ac} = R_{dc} = .0085\Omega$$

10. Calculate winding loss, total loss, and temperature rise, *using the interleaved structure*:

Secondary dc loss (R_{dc} = .00049Ω):

$$P_{Sdc} = 12^2 \cdot .00049 = 0.07 \text{ Watts}$$

Secondary ac loss (R_{ac} = .0007Ω):

$$P_{Sac} = I_{Sac}{}^2 \cdot R_{ac} = 15^2 \cdot .0007 = 0.16 \text{ W}$$

Total secondary winding loss -- dc plus ac is:

$$P_{Sw} = 0.07 + 0.16 = 0.23 \text{ Watts}$$

Primary dc loss (R_{dc} = .0085Ω):

$$P_{Pdc} = I_{Pdc}{}^2 R_{dc} = 2.8^2 \times .0085 = .067 \text{ Watts}$$

Primary ac loss (R_{ac} = .0085Ω):

$$P_{Pac} = I_{Pac}{}^2 R_{ac} = 3.71^2 \times .0085 = 0.12 \text{ Watts}$$

Total primary winding loss -- dc plus ac is:

$$P_{Pw} = .07 + 0.12 = 0.19 \text{ Watts}$$

Total winding loss:

$$P_W = 0.23 + 0.19 = 0.42 \text{ Watts}$$

Total loss, including core loss of 0.56 W:

$$P_T = P_W + P_C = 0.42 + 0.56 = 0.98 \text{ W}$$

The total loss is well within the 1.42 Watt limit. The temperature rise will be:

$$^\circ Crise = R_T \times P_T = 28 \times 0.98 = 27\ ^\circ C$$

Total winding height, including two layers of .02cm isolation: 0.04 + 0.081 + .112 = **0.233 cm**, well within the 0.38 cm available.

Leakage inductance between primary and secondary calculated according to the procedure presented in Reference R3 is approximately .08μH, referred to the primary. Interwinding capacitance is approximately 50pF.

If the windings were not interleaved (as shown in Figure 5-8), leakage inductance would be more than doubled, but interwinding capacitance would be halved. Interleaving also helps to reduce ac winding losses. This becomes much more significant at higher power levels with larger conductor sizes.

It is very important to minimize leakage inductance in flyback circuits. Leakage inductance not only slows down switching transitions and dumps its energy into a clamp, but it can steal a very significant amount of energy from the mutual inductance and prevent it from being delivered to the output.

References

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

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

(R5) "Coupled Filter Inductors in Multiple Output Buck Regulators Provide Dramatic Performance Improvement," *Unitrode Seminar Manual SEM1100,* 1996

(1) MIT Staff, "Magnetic Circuits and Transformers," *MIT Press, 1943*

Reference Design Section