Section 3 Windings

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

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

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

Conservation of Energy

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

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

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

Skin Effect

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

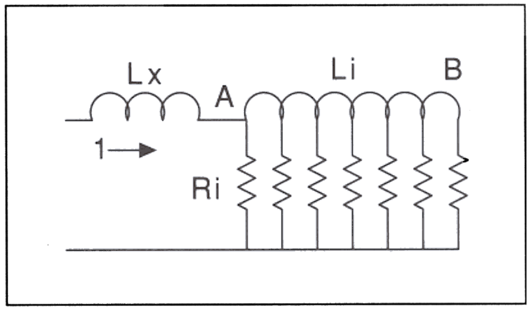


Figure 3-1 Skin Effect Model

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

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

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

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

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

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

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

Proximity Effect

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

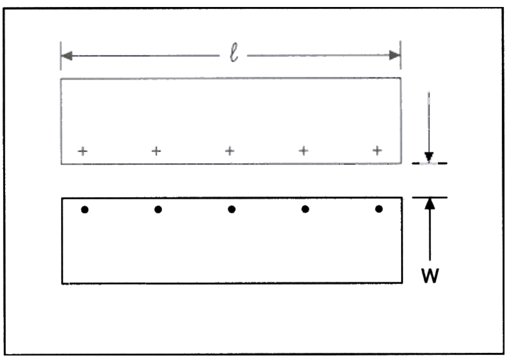


Fig. 3-2 Proximity Effect

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

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

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

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

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

Examples

In the examples of winding structures given below:

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

Simple Transformer Windings:

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

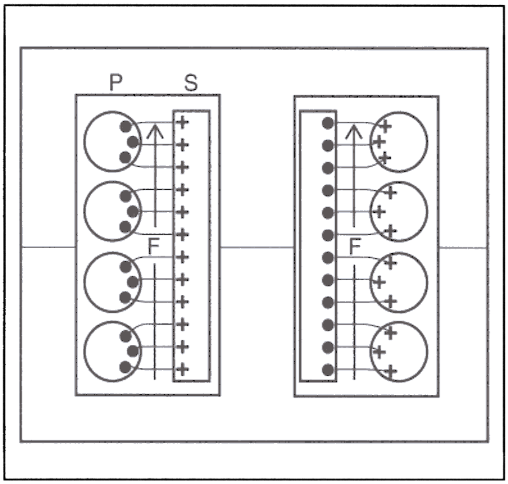


Figure 3-3 Single Layer Windings

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

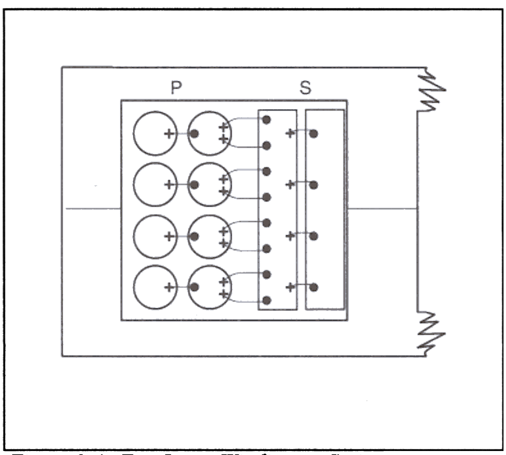


Figure 3-4 Two Layer Windings -- Series

Multiple Layer Windings:

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

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

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

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

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

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

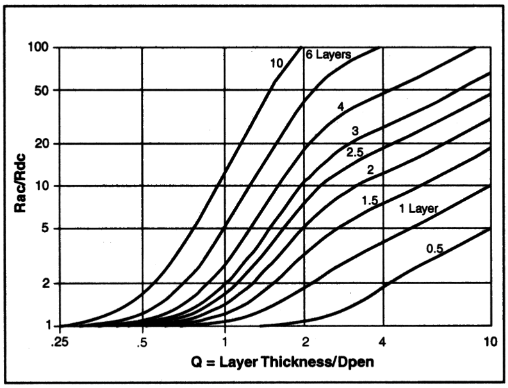


Figure. 3-5 Eddy Current Losses -- R_{AC}/R_{DC}

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

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

Alternative methods of calculating eddy current losses include:

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

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

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

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

Paralleled Layers:

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

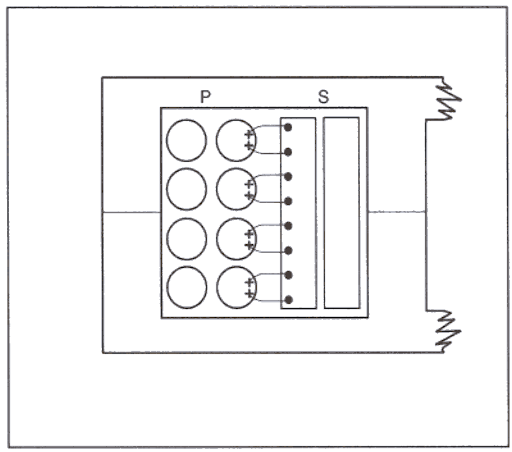


Figure 3-6 Paralleled Two-Layer Windings

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

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

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

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

Window Shape–Maximize Breadth

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

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

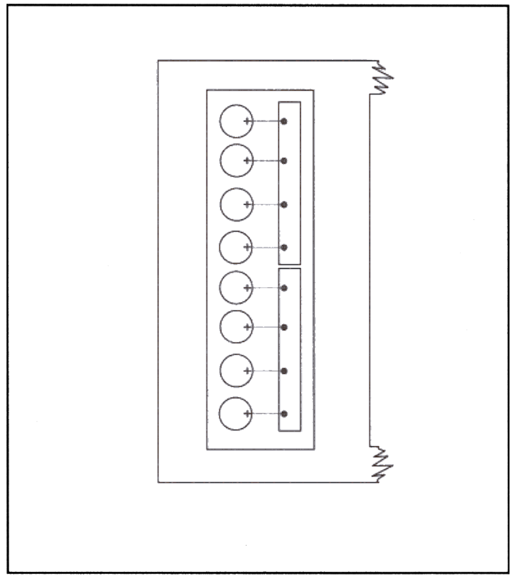


Figure. 3-7 Wide Window Breadth

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

Interleaving:

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

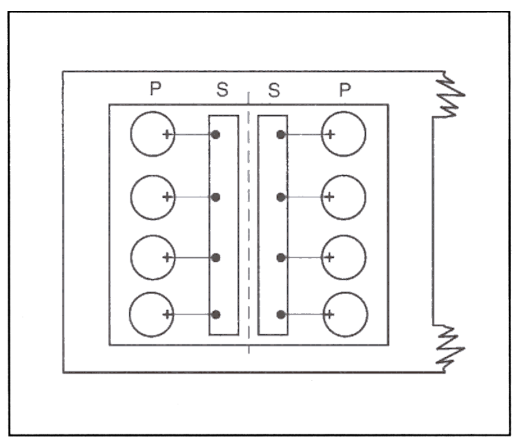


Figure. 3-8 Interleaved Windings

Winding Sections:

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

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

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

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

When does Paralleling Succeed:

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

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

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

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

Passive losses

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

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

Passive winding losses can be reduced or eliminated by:

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

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

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

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

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

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

Calculating rms Current

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

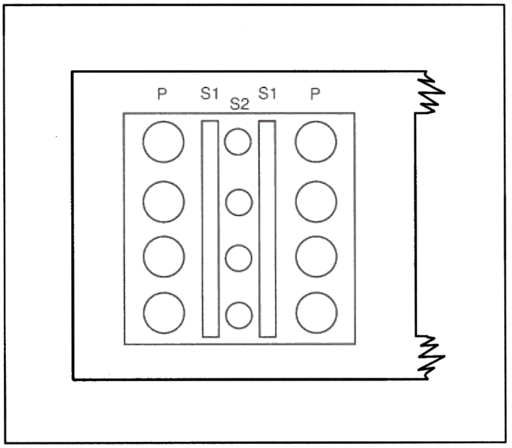
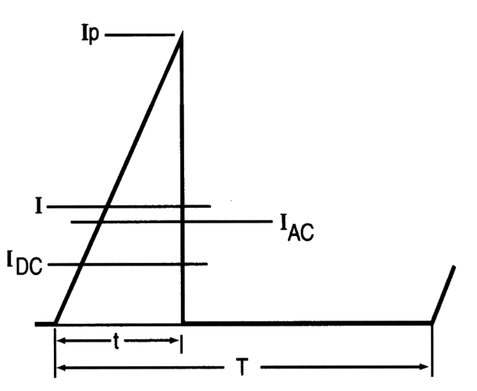


Figure 3-9 Interleaved Winding Hierarchy

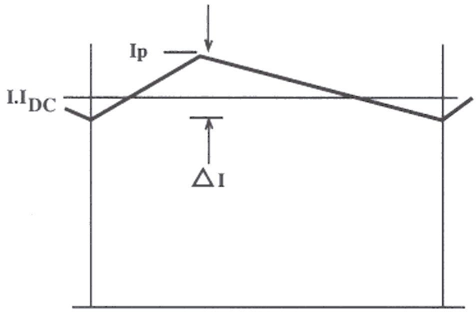
shapes encountered in switching power supplies. For each half of a Center-Tap winding, which conducts only on alternate switching periods, substitute D/2 in place of D.

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



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

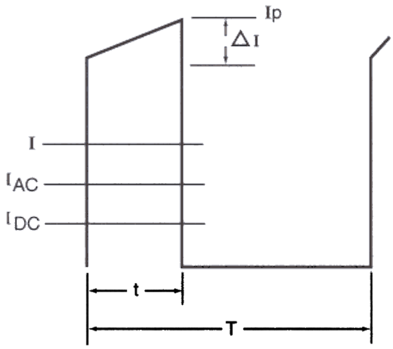
Fig 3-10a Discontinuous Mode Waveform



$$I \approx I_{DC} = I_P - \Delta I/2$$

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

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



$$I_{DC} = (I_P - \Delta I/2)D$$

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

$$I = (I_P - \Delta I/2)D^{1/2}$$

Figure. 3-10c Continuous Mode -- Transformer

Window Utilization

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

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

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

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

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

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

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

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

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

Triple-Insulated Wire

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

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

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

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

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

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

Optimized Winding Placement

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

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

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

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

Inter-Winding Capacitance

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

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

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

End-to-end capacitance

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

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

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

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

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

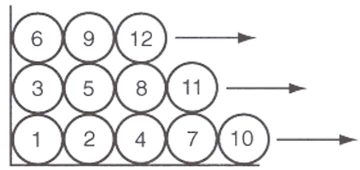


Figure. 3-11 Bank Winding

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

Points to Remember

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

causing losses to rise exponentially with the number of layers.

Know the rules, and make them work for you, rather than against you.

Using curves derived from Dowell, subdivide conductors into smaller diameter paralleled wires (Litz) to achieve R_{AC}/R_{DC} of 1.5. But a point may be reached where increased voids and insulation cause R_{DC} to rise to the point where further subdivision is unproductive.

Leakage inductance energy must be absorbed by snubbers or clamps, usually resulting in load dependent losses. Leakage inductance is also the main cause of impaired cross-regulation between multiple outputs.

Minimize leakage inductance *and* eddy current loss by using a window shape that maximizes winding breadth, and/or by interleaving the windings. There is a penalty in increased interwinding capacitance.

Leakage inductance is increased by physical separation mandated by agency high voltage isolation requirements. Consider using triple-insulated wire, especially in low-power applications where isolation penalty is more severe.

Distribute transformer windings so as to intimately conform to each other, to minimize leakage inductance and external stray fields. Place inductor windings to conform to the gap as closely as possible, whether a distributed gap or a discrete gap, to prevent build-up of magnetic force which propagates stray field.

Minimize the *effect* of leakage inductance by using the appropriate winding hierarchy–the highest power secondary should be closest coupled to the primary. Keep the lower power windings out high field intensity region between primary and high-power secondaries.

Avoid center-tapped primary windings. It would be nice if center-tapped secondaries could be avoided, as well.

References:

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(3) K. O'Meara, "Proximity Losses in AC Magnetic Devices," *PCIM Magazine,* December 1996

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

(5) B. Carsten, "Switchmode Design and Layout Techniques," APEC'97 Tutorial

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