

# Designing Planar Magnetics

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## ABSTRACT

*Planar magnetic devices offer several advantages over their conventional counterparts. This paper discusses the magnetic fields within the planar structure and their effects on the distribution of high frequency currents in the windings. Strategies for optimizing planar design are presented, and illustrated with design examples. Circuit topologies best suited for high frequency applications are discussed.*

## I. ADVANTAGES OF PLANAR MAGNETICS

In contrast to the helical windings of conventional magnetic devices, the windings of planar transformers and inductors are located on flat surfaces extending outward from the core centerleg.



Fig. 1. Planar transformer.

Magnetic cores used with planar devices have a different shape than conventional cores used with helical windings. Compared to a conventional magnetic core of equal core volume, devices built with optimized planar magnetic cores usually exhibit:

- Significantly reduced height (low profile)
- Greater surface area, resulting in improved heat dissipation capability.
- Greater magnetic cross-section area, enabling fewer turns
- Smaller winding area
- Winding structure facilitates interleaving
- Lower leakage inductance resulting from fewer turns and interleaved windings
- Less AC winding resistance
- Excellent reproducibility, enabled by winding structure

In transformer applications, the winding configurations commonly employed in planar devices are advantageous in reducing AC winding losses. However, in inductors and flyback transformers using gapped centerlegs, the winding configuration often results in greater AC winding loss.

## II. MAGNETIC FIELD PROPERTIES

A magnetic field is actually stored energy. The physical distribution of the magnetic field represents the distribution of this energy. Understanding the properties of the magnetic field not only reveals the amount of stored energy and its locations, it also reveals how and where this energy is coupled to various electrical circuit elements.

Magnetic / electric relationships are much simpler and easier to understand when using the SI system of units (rationalized MKS). When core and winding materials are specified in CGS or English units, it is nevertheless best to think in SI units throughout the design process, and then if necessary convert to other unit systems as the final step in the process.

Tutorials on magnetics design have been presented at previous Unitrode/TI seminars. Most of this material has been consolidated into a "Magnetics Design Handbook, MAG100A". This handbook, as well as all past seminar topics, is available for downloading from the web site <http://power.ti.com>. Click on [Design Resources] → [Power Management Training].

*Inductance* is simply an electrical circuit concept which enables the circuit designer to predict and quantify the effects of magnetically stored energy in the electrical circuit.

Applying the basic principles of magnetic field behavior (discussed in earlier seminars) to planar magnetic structures enables us to optimize the design and predict the magnitude of parasitic circuit elements such as leakage inductance. The magnetic field also is the dominant influence on the distribution of high frequency AC current in the windings, thereby determining AC winding losses.

### A. Review of Magnetic Field Fundamentals

Rules governing magnetic field behavior are summarized in Appendix I.

Every magnetic field has two components: *Magnetic force*,  $F$ , (magnetic potential), and *magnetic flux*,  $\Phi$ . Magnetic force is directly proportional to current (Ampere’s Law). In fact, in the SI system of units, magnetic force,  $F$ , *directly equals* current – units of magnetic force are expressed in Amperes. Thus, 1 Ampere of current flowing in a conductor inevitably results in 1 Ampere of magnetic force.

Magnetic force can be described as a series of equipotential surfaces. The spacing of these surfaces defines a force gradient – a magnetic potential gradient. The magnitude of this gradient at any location is called *field intensity*,  $H$ .

Fig. 2. shows the leakage inductance in the left half of a planar transformer structure. (In order to provide clarity of illustration, only three primary turns are used, and spacing between primary and secondary is greatly exaggerated.) The light dash lines show the edge view of the magnetic force equipotentials between primary and secondary windings. The light solid lines represent flux. The equipotential surfaces can be thought of as elastic membranes which terminate on current flow and are “anchored” on the opposing currents which produce the field.

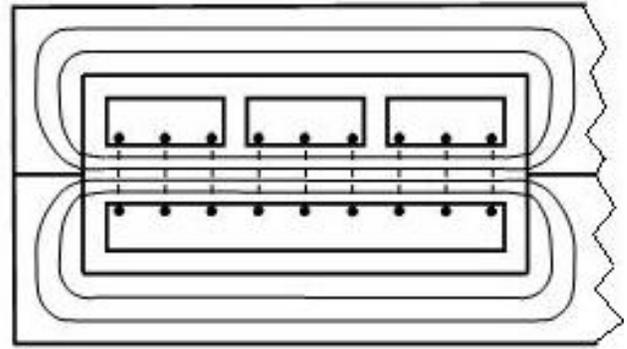


Fig. 2. Cross-section of equipotentials and flux lines within a planar transformer (one-half of transformer shown).

DC and AC current distributions within the windings usually differ significantly. At high frequencies, the magnetic field arranges itself so as to minimize the rate of energy transfer between the electrical circuit and the field. The field “pulls” the opposing currents to the conductor surfaces closest to each other, as shown in Fig. 2., thereby minimizing the volume of the field (skin/proximity effect). Also, the currents spread across the opposing conductor surfaces so as to minimize energy density.

However, at low frequencies, the rate of energy transfer between the circuit and the magnetic field is very small. The rate of energy transfer into the conductor resistance is greater. Therefore, DC and low frequency currents distribute uniformly throughout the conductors so as to minimize  $I^2R$  loss.

### III. THE “TRUE” TRANSFORMER

Transformers in switching power supplies are used primarily in buck-derived topologies (forward converter, full bridge, half bridge, etc.) In a transformer, energy storage is usually undesirable, but unavoidable – appearing in the transformer equivalent circuit as parasitic leakage inductance and magnetizing inductance. (Flyback transformers are misnamed – they are actually coupled inductors. Energy storage is essential to their function.)

In a “true” transformer (Fig. 2.), opposing currents flow simultaneously in primary and secondary windings. The Ampere-turns in the secondary winding, resulting from load current, are canceled by equal and opposite Ampere-turns

flowing in the primary. A small additional unopposed magnetizing current also flows in the windings. This magnetizing current provides the small magnetic force necessary to push flux through the very low reluctance of the high permeability magnetic core. The closed loops of this magnetizing flux link the primary and secondary windings to each other, thus providing the coupling which is essential for transformer operation (shown in Fig. 3.). Magnetizing flux and its associated magnetizing current change as a function of Volt-seconds per turn applied to the windings (Faraday's Law) *independently of load current*. Magnetizing inductance appears in the transformer equivalent electrical circuit as a shunt element.

Much of the energy stored in the magnetizing inductance goes into hysteresis loss, the rest is usually dissipated in snubbers or clamps. If the core were ideal – with infinite permeability – the magnetizing inductance value would be infinite, and thus have no effect on circuit performance.

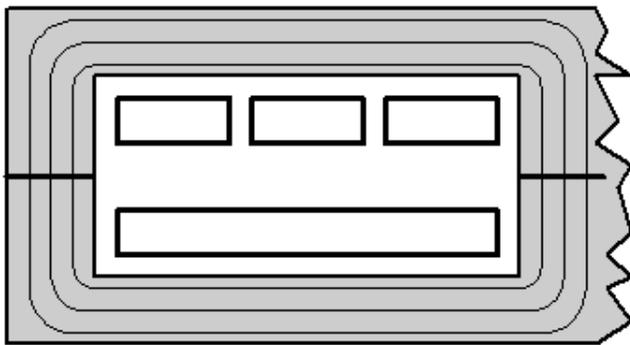


Fig. 3. Magnetizing flux links the windings.

Excluding magnetizing current, load-related Ampere-turns in primary and secondary windings cancel completely. Load current has no effect on the magnetizing flux in the core. Magnetic force related to load current exists in only one place within the transformer – in the region *between* primary and secondary windings where the currents do not cancel. As shown in Fig. 2., the flux lines associated with this field between the windings link half the energy of the field to primary and half to the secondary winding. But these flux lines do not link the windings to each other. Thus, the coupling between windings is impaired. The energy stored in this inter-winding region appears in the equivalent electrical circuit as se-

ries *leakage inductances*, ( $L_{LP}$ ,  $L_{LS}$ ). Each time the power switch turns off, energy stored in the leakage inductance usually ends up dissipated in snubbers or clamps, thus degrading power supply efficiency.

Fig. 4. shows the electrical equivalent circuit of the transformer, including the magnetizing inductance and parasitic leakage inductance apportioned to primary and secondary windings. The “ideal transformer” is used to account for the actual turns ratio and primary-secondary isolation. Leakage inductances are usually so small compared to the magnetizing inductance value that they can be combined, with negligible error, into a single leakage inductance value in an equivalent “L” network. Magnetizing inductance can be assigned to either the primary or secondary side.

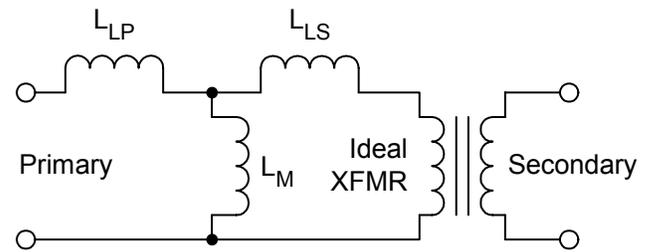


Fig. 4. Transformer equivalent electrical circuit.

The leakage inductance value can be calculated from the physical dimensions of the windings<sup>[1]</sup>. Leakage inductance is minimized by:

- minimizing the number of turns
- using a core with a large winding “breadth”
- interleaving windings
- minimizing the spacing between primary and secondary windings

Bifilar windings approach the ideal, but this is usually not possible in a planar transformer, especially when high voltage isolation is required.

### A. Current Flow Patterns -- Transformer

Fig. 5. shows the waveforms of the Ampere-turns in the transformer windings (Ampere-turns are identified with the symbol  $I'$ ). Load-related current flows simultaneously in primary and secondary windings. The combined primary and secondary Ampere-turns are equal and opposite, thus canceling each other except for the magnetizing current (dash lines) which provides the small magnetic force required to push the coupling flux through the core. The differential Ampere-turns provide the much larger force between the primary and secondary, creating the undesired leakage inductance field.

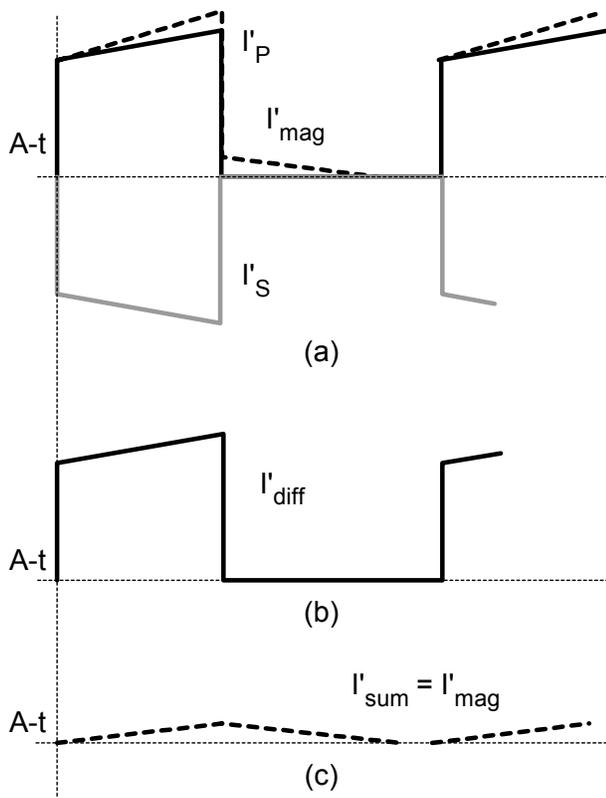


Fig. 5. Forward converter waveforms.

The high frequency magnetic field equipotential surfaces, shown in Fig. 2., “pull” the opposing high frequency current components to the closest conductor surfaces in order to minimize the volume of the magnetic field, thereby minimizing the stored energy. The high frequency component of the secondary current is distributed uniformly across the full width of the single turn secondary, directly opposite the opposing pri-

mary current, in order to minimize energy density.

The winding geometry of the planar transformer is inherently favorable because high frequency currents are spread across the broad conductor surfaces facing each other, thus minimizing high frequency losses.

If the conductor thickness is greater than the high frequency skin depth, high frequency current flow will be concentrated within one skin depth of the surfaces directly facing each other. Little or no current will flow on the outer surfaces of these conductors.

DC current distribution differs significantly from the high frequency AC distribution. Low frequency distribution is determined entirely by the resistance of the conductors.

The DC *primary* current component is distributed uniformly across the winding breadth. It must, because the primary turns across the breadth are in series. However, in the single turn secondary, the DC current component concentrates somewhat toward the center, because the distance around the turn is shorter and the resistance less. The DC magnetic field is skewed and not minimized. But the field does not dictate DC current distribution because the rate of energy transfer in and out of the DC field is zero. Resistance rules!

Skin depth for DC current is infinite, so the DC current components are distributed uniformly throughout the thickness of the conductor. In the structure of Fig. 2., with single layer primary and secondary, the conductors could be made much thicker than the high frequency skin depth. This would reduce the DC resistance, but would not reduce high frequency AC resistance.

**Warning:** If windings have multiple layers, conductor thickness greater than the skin depth will actually result in dramatically *increased* high frequency losses. If multiple layers are required, primary and secondary conductors must be much thinner than the skin depth, or, primary and secondary layers must be interleaved, forming winding sections, with one layer in each section. This is discussed in detail in [2].

#### IV. INDUCTORS AND FLYBACK TRANSFORMERS

In a true transformer, inductance is undesirable. In an inductor or flyback “transformer” (which is really an inductor with multiple windings), inductive energy storage is not only desirable, it is essential. However, very little energy can be stored within the high permeability magnetic core material, and most of the energy that is stored in the core is unrecoverable, ending up as core loss or dissipated in snubbers or clamps. In order to efficiently store the energy required in an inductor or flyback transformer, a non-magnetic “gap” (often an air gap) is placed in series with the flux path within the core, usually in the core centerleg.

In a flyback transformer, primary and secondary currents occur *sequentially*, not simultaneously as in a true transformer. Therefore they do not cancel. Current in the flyback transformer windings is *entirely magnetizing current*, with the associated magnetic force appearing almost entirely across the high reluctance non-magnetic gap where the energy is stored. However, in a manner similar to a true transformer, parasitic leakage inductance between primary and secondary windings does impede current transfer between primary and secondary when the power switch turns on and off.

The field storing the desired energy (in the gap) is created by the *combined* Ampere-turns of the windings. The undesired leakage inductance field is created by the *differential* Ampere-turns between primary and secondary windings. Thus, in order to define the magnetic field configurations and the resulting current distributions within the windings, it is necessary to define the current components.

#### V. CONTINUOUS MODE FLYBACK TRANSFORMER

The first example is a flyback transformer operating in the continuous inductor current mode (the combined currents in the windings never dwell at zero). As shown in Fig. 6., the current in the 3-turn primary is 1 A, with a duty cycle,  $D$ , of 0.33.

As shown in Fig. 6a., when the primary power switch turns off, the 3 A-t primary current transfers to the secondary which then has 3 A in its single turn. (The transitions are not instantaneous because of leakage inductance.) The primary and secondary Ampere-turns combine sequentially to produce a continuous 3 Ampere field within the centerleg gap – the *gap field* – which is mostly DC. The inductance value is large, so the ripple component of the combined Ampere-turns is small. In the same proportion, the AC field component is small compared to the DC component, so that flux density in the core is more likely limited by saturation rather than by core loss.

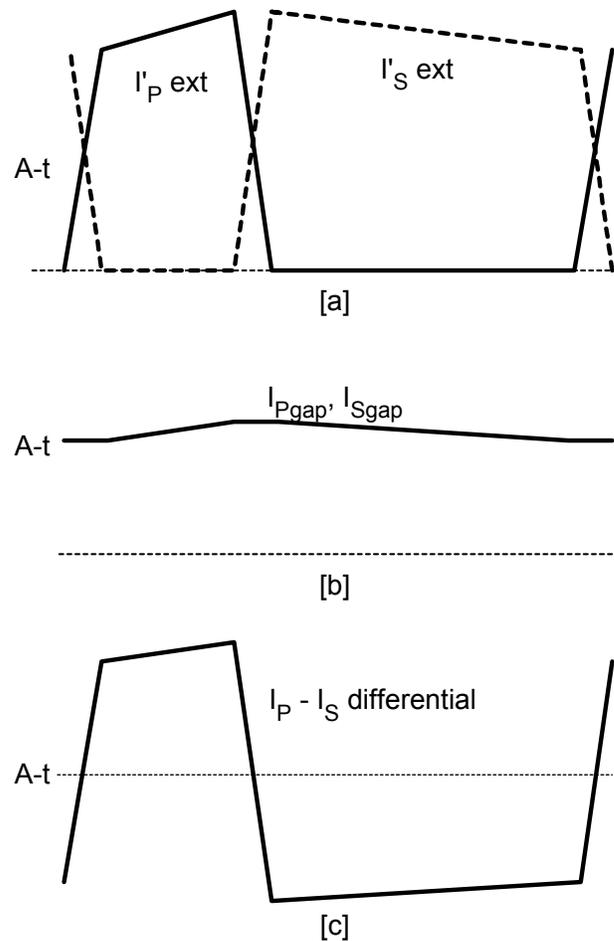


Fig. 6. Continuous mode flyback waveforms.

The current flow and magnetic field distributions can be difficult to explain and to understand. The field expands to fill the entire winding region in order to provide flux continuity across this region. Although the current alternates from

one winding to the other, the resulting field is relatively constant in the gap and all regions outside the gap, *except* in the region between the windings. Here, there is a field component which changes direction as current alternates between primary and secondary. The energy associated with this changing field is leakage inductance energy. Its reversal requires energy which impedes current transfer, just as in the true transformer discussed previously.

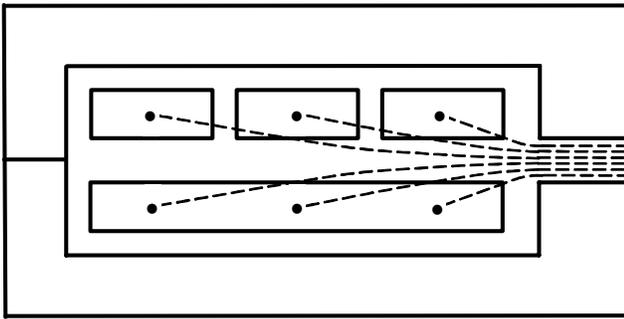


Fig. 7. Flyback transformer with DC field pattern through gap (one-half shown).

In order to better understand and to quantify the current flow patterns within the windings and the resulting magnetic field distributions, it is helpful, using superposition, to break the field down into several components.

Remember that every magnetic force equipotential must be terminated by equal and opposite Ampere-turns. The gap field is terminated all around its periphery by the currents in the windings flowing around the outside of the gap. The *leakage field* between the primary and secondary windings requires equal and opposite A-t flowing simultaneously in these two windings. However, currents are applied *sequentially* in primary and secondary windings, so that the external current applied to one or the other of these two windings is always zero. How, then, can the leakage field between the windings be terminated properly?

Because of the symmetry of the windings with respect to the gap, the primary and secondary windings each provide approximately *one-half* of the *combined* Ampere-turns driving the gap field. As shown in Fig. 6b. and Fig. 7., each winding supplies half of the combined A-t *continuously*, regardless of duty cycle – *even when no current is flowing externally in that winding*.

For example, at the time when 1 A is being applied to the 3-turn primary winding (3 A-t), half of this, 1.5 A-t, is contributed directly from the primary to the gap field. The remaining primary 1.5 A-t initiates one side of the leakage inductance field between the primary and secondary. The other side of the leakage field *must* be terminated by *minus* 1.5 A-t in the secondary winding. That *minus* 1.5 A-t balances out the *plus* 1.5 A-t that the secondary is providing continuously to the gap field, even though at that moment, the external secondary current is zero.

Later in the switching period, when the external secondary current is 3 A and external primary current is zero, the secondary provides *plus* 1.5 A-t directly to the gap field, and *plus* 1.5 A-t to the leakage inductance field. The primary then has *minus* 1.5 A-t to terminate the leakage inductance field which balances its continuous *plus* 1.5 A-t contribution to the gap field.

Thus, although the currents applied to the windings are sequential, each winding contributes *continuously* its half share of the total magnetic force within the gap. These currents driving the gap field are essentially DC, and distribute themselves accordingly within the windings so as to minimize  $I^2R$  losses.

The leakage inductance field (Fig. 6c.) alternates its direction between the windings. This leakage inductance field and the currents driving it have a very large AC component, but – as with the “true” transformer – the leakage inductance field pulls these opposing AC currents to the broad surfaces of the conductors, thereby minimizing AC losses.

But the AC component of the gap field pulls the AC component of the combined Ampere-turns to the narrow edges of the conductors facing the gap. The result is extremely high AC resistance, but because the AC component is small in the continuous mode, the AC losses may be acceptable.

Thus, there are several current components within each winding which are separate and distinct, *distributed differently* and are never zero, even though they cancel during the times when external currents are zero.

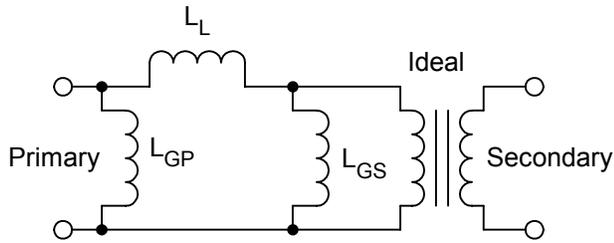


Fig. 8. Flyback transformer equivalent circuit.

The equivalent electrical circuit for the flyback transformer is shown in Fig. 8. The shunt mutual inductances in the flyback transformer on the primary and secondary side together represent the energy stored in the gap. Note that because the leakage inductance value is relatively small, continuous current flows in both the primary and secondary side mutual inductances, regardless of which winding has external current flow. This equivalent circuit may help to explain the continuous primary and secondary current components driving the gap field, as shown in Fig. 6b.

## VI. DISCONTINUOUS MODE FLYBACK TRANSFORMER

With discontinuous mode operation, when the primary power switch turns on, primary current starts at zero and rises to a peak value – at least twice the peak current in a comparable continuous mode application. When the power switch turns off, the primary Ampere-turns transfer to the secondary (although impeded by leakage inductance). Secondary current then decreases to zero and remains at zero (by definition, for discontinuous mode operation) until the beginning of the next switching period.

For discontinuous mode operation, the transformer is designed with a much smaller inductance value. Although the physical structure is the same as the continuous mode flyback, Fig. 6., the core size is probably smaller and the gap size and the numbers of turns would differ, for the same power supply application.

Fig. 9. shows the worst case for discontinuous mode transformer operation, at the mode boundary between discontinuous and continuous, where the rms currents are greatest.

As with continuous mode operation, the combined sequential primary and secondary Ampere-turns create the magnetic field in the gap. But unlike the continuous mode, there is a very large AC component of the gap field (Fig. 9b.), and of the flux within the gap and the core. Because of the large AC component, flux density is almost certainly limited by core loss.

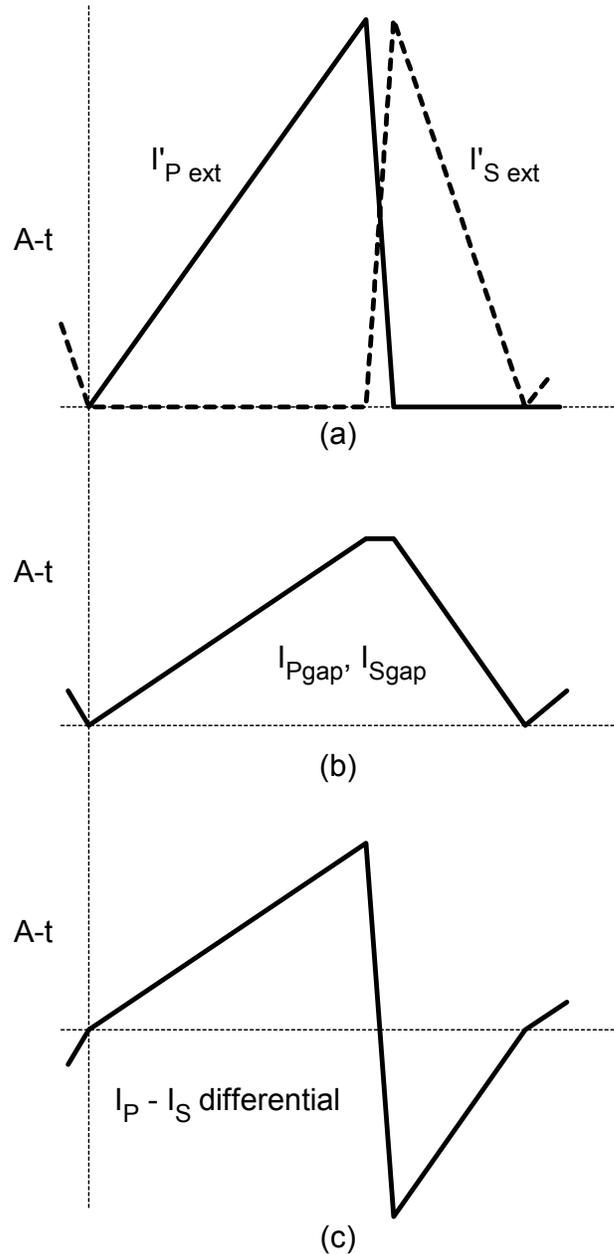


Fig. 9. Flyback transformer currents, discontinuous mode, at mode boundary.

As shown in Fig. 9b, the primary and secondary each provide one-half the combined magnetic force, simultaneously, and not sequentially. Also, both windings provide the currents to properly terminate the leakage inductance field (Fig. 9c.), as with continuous mode operation. The currents terminating the leakage field are pulled to the broad conductor surfaces, minimizing winding losses due to the leakage inductance field.

The important difference with continuous mode operation is that discontinuous operation results in a very large AC component in the gap field.

### A. Current is Pulled Toward the Gap

In a conventional core with helical, cylindrical windings, the best place to locate the required gap is in the core centerleg. However, in a planar core, a centerleg gap results in the magnetic force equipotentials (which pass through the gap from one side of the winding to the other) “pulling” the high frequency AC current toward the narrow edges of the conductors facing the gap, as shown in Fig. 10. (Compare this with the DC current and field distributions in Fig. 7.) This “edge effect” results in very high AC winding resistance. DC current components will distribute themselves according to conductor resistances, because the magnetic field does not influence DC current distributions.

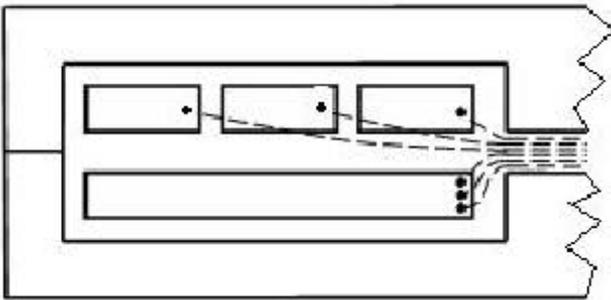


Fig. 10. AC current pulled toward gap.

The high AC resistance due to the edge effect may be acceptable in a filter inductor or continuous mode flyback application where the AC field component in the gap is usually much smaller than the DC component. But in a discontinuous mode flyback transformer, where the AC component is relatively large, AC losses easily become unacceptable – even catastrophic.

This problem can be avoided by relocating the gap to a position where the magnetic force equipotentials pull the current to the broad surfaces of the conductors, rather than to the narrow edges. Locating a gap in any of the outside surfaces of the core will not work, because to create a field within the gap would then require opposing conductors outside of the gap, i.e., outside of the entire core. Where, then, can the gap be placed?

### B. A Better Gap Location

A solution to this problem is to make the height of the winding window somewhat greater than required to contain the winding, as shown in Fig. 11. The winding should be placed at one extremity of the available window height, and the gap placed at the opposite extremity. The field *must* distribute itself rather uniformly throughout the empty portion of the window, in order to minimize stored energy, and to provide the necessary continuity of the flux lines through this region (flux lines can never terminate). Thus, the force equipotentials advantageously “pull” the current to the broad conductor surfaces, rather than to the narrow edges.

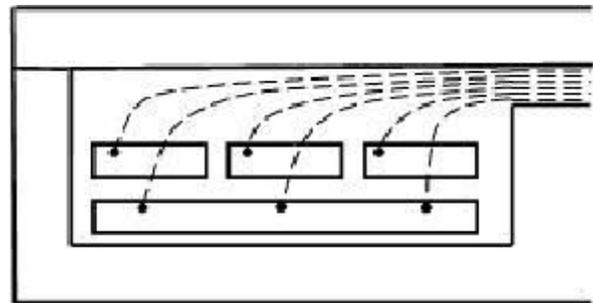


Fig. 11. Gap relocated.

Energy stored in this empty portion of the window contributes to the inductance value. It may be significant, but difficult to calculate. This will require adjustment of the centerleg gap in order to achieve the desired inductance value.

A disadvantage of this technique is that it requires some increase in overall core height.

The conductor furthest from the gap (the secondary in Fig. 11) can be much thicker than the skin depth, reducing DC loss and assisting heat removal. But although the high frequency AC currents spread across the broad surfaces of the conductors rather than the edges – a definite advantage – the gap field passes through all other windings closer to the gap. This requires these closer windings to be restricted in thickness to approximately the skin depth, to avoid large AC losses. Interleaving the windings has the same advantage as in the “true” transformer (see Fig. 16.) in reducing leakage inductance, but because of the direction of the gap field, *through* the windings, all of the interleaved windings must be thin except the winding furthest from the gap.

### VII. CORE SHAPE

A good core shape for a planar transformer is shown in Fig. 12. The magnetic cross section area should be large in order to minimize the number of turns in the windings. Winding breadth “ $b_w$ ” should be as large as possible in order to minimize leakage inductance and to maximize the width of the copper traces (which are limited in effective thickness by skin depth and/or PC board plating considerations). The core should cover the winding as much as possible. This minimizes stray field and EMI, improves heat dissipation, and permits a small reduction of core height. A round centerleg minimizes the length of winding turns, hence reduces copper losses and leakage inductance.

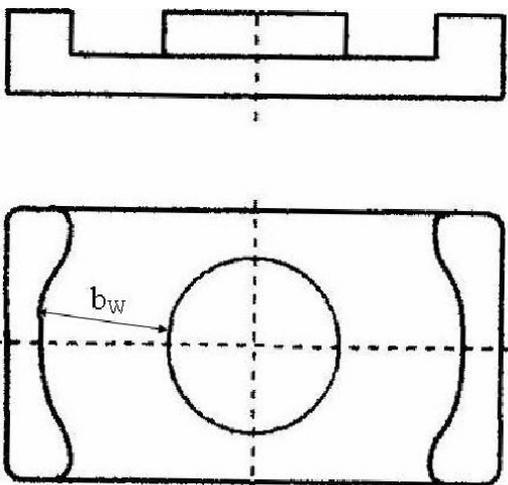


Fig. 12. Planar core shape.

The core shape shown in Fig. 13. – a modified E-E or E-I core shape – is easier for the core manufacturer to fabricate, but the elongated rectangular centerleg results in increased winding length. However, greater winding breadth is an important advantage. Since much of the winding is exposed, stray field and EMI are greater.

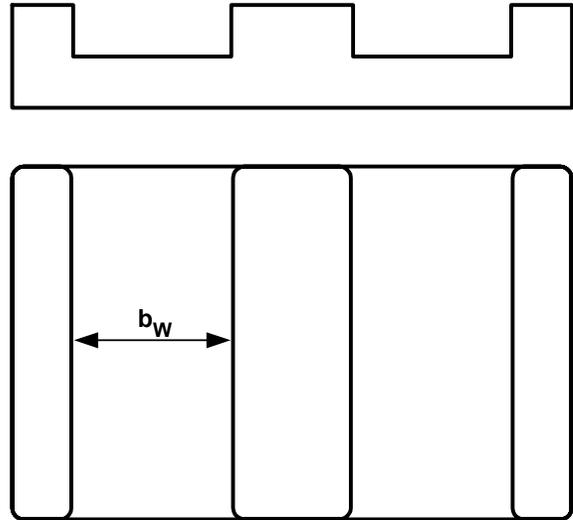


Fig. 13. Core with rectangular centerleg.

### VIII. CURVED VS. STRAIGHT WINDING PORTIONS

On a core with round centerleg, as shown in Fig. 12., windings are circular throughout. The leakage inductance field between the windings is radial. On a core with a rectangular centerleg as shown in Fig. 13., windings have linear portions parallel to the centerleg, and have curved portions around the ends.

Fig. 14. shows a cross-section of an interwinding region where the turns are straight and parallel to each other. The field between the windings is leakage inductance energy. Since the current in the windings and resulting magnetic force equipotential surfaces are evenly spaced, the flux density throughout this region is constant. The same number of flux lines exist throughout the region. Flux lines are always closed loops, encircling the current that is producing the field. As shown in Fig. 14., because of inherent symmetry, half of the flux lines encircle the upper winding and complete their paths through the upper portion of the core, while the other half of the flux lines encircle the lower

winding. Thus, the flux lines link the leakage inductance energy to all the turns of either the upper or the lower winding, but none of these flux lines link the upper and lower windings to each other.

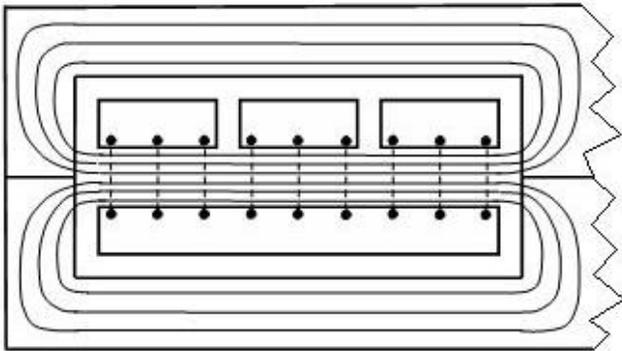


Fig. 14. Leakage field, straight winding portion.

Fig. 15. shows a similar cross-section of an inter-winding region where the turns are curved, such as when encircling a round centerleg. (The curvature cannot be seen because it is in a plane normal to the view plane.) The flux lines shown represent the flux in an arc segment – like looking at the edge view of a piece of pie.

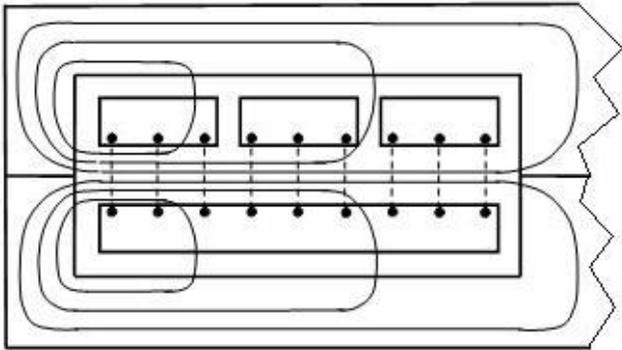


Fig. 15. Leakage field, curved winding portion.

Thus, the length of each turn segment is proportional to its distance from the center of curvature. Nevertheless, because the turns are equally spaced, the field intensity, and therefore the flux density *and the energy density* are all constant throughout the region. But because the flux density is constant, the total flux in the outer regions of the curved windings is proportionately greater than the total flux in the inner regions. Since all flux lines must be closed loops, is not possible for all flux lines to link all turns of either winding. Some of the flux lines in the outer region link only to the outer turns, as shown in Fig. 15.

Flux lines reveal how the energy of the magnetic field is coupled to the windings. The flux pattern of Fig. 15 shows that more energy is coupled to the outer turns than to the inner turns. This is consistent with the fact that the outer turns are longer, thus encompassing more of the constant energy density leakage inductance field, therefore contributing more to the leakage inductance value.

Regardless of the manner in which the energy is coupled, all of the inter-winding energy is coupled to the windings on either side, and all of this inter-winding energy shows up in the electrical equivalent circuit as energy stored in the leakage inductance.

*To summarize:* Regardless of the curvature of the windings and the configuration of the flux lines, *all* of the energy in the leakage inductance field is coupled to either the upper or the lower windings, as shown in Fig. 14. and Fig. 15.

Regardless of curvature, if the windings are equally spaced, energy density is uniform in the inter-winding region, and can be calculated from the Ampere-turns distributed across the winding breadth. Total leakage inductance energy can then be calculated from energy density multiplied by total inter-winding volume.

## IX. WINDING PORTIONS EXTERNAL TO THE CORE

In an “old style” transformer with helical windings around a central core, the ends of the helical windings are close to the core, even where the windings emerge from the core, so almost all of the leakage flux lines take their return path through the low reluctance core.

In a planar transformer using a core shape such as that of Fig. 12., the windings are almost totally enclosed within the core, so that the leakage flux has an easy return path through the core. But with cores such as in Fig. 13., a substantial portion of the windings extend beyond the core. Leakage flux lines have a much longer path to reach the core, with the result that many extend outward beyond the core, thus creating EMI.

## X. CALCULATING AND MINIMIZING LEAKAGE INDUCTANCE

Applying the few simple rules of magnetic field behavior, the configuration and magnitude of the leakage inductance field between windings is easily calculated.

According to Ampere's Law, the total magnetic force across the inter-winding region (the leakage inductance field) is equal to the load-related Ampere-turns flowing in the windings. If the windings are spaced uniformly across the winding breadth, the field intensity,  $H$ , indicated by the spacing between equipotential surfaces, is constant throughout the inter-winding region. Field intensity,  $H$ , equals  $F$  divided by winding breadth,  $b_w$ . Hence,

$$H = \frac{F}{b_w} = \frac{NI}{b_w} \quad (1)$$

The permeability in this "nonmagnetic" region between the windings is constant ( $\mu_0 = 4\pi \cdot 10^{-7}$ ), hence the flux density,  $B$ , is also constant ( $B = \mu H$ ).

In any region where permeability is constant, energy density equals  $\frac{1}{2}BH$ . Thus, the energy density,  $E/v$ , throughout the leakage inductance field between windings is fairly constant.

$$E/v = \frac{1}{2}BH = \frac{1}{2}\mu_0 H^2 = \frac{1}{2}\mu_0 \left(\frac{NI}{b_w}\right)^2 \text{ Joules/m}^3 \quad (2)$$

Total energy,  $E$ , can then be calculated by multiplying energy density times the total volume of the inter-winding region.

The leakage inductance value, seen from the perspective of any winding, can then be calculated:

$$E = \frac{1}{2}LI^2; \quad \therefore L_L = 2E/I^2 \quad (3)$$

where  $I$  is the current in the referenced winding. Leakage inductance is minimized by increasing winding breadth  $b_w$ , minimizing the number of turns, spacing windings as close as possible, reducing the length of turns (a round centerleg is best) and by interleaving the windings.

## XI. WINDING CONSIDERATIONS

The windings can easily be the greatest design challenge in a high frequency planar transformer. Reference [2] should be consulted to obtain the necessary understanding for design of the windings.

High-frequency current is restricted to the conductor surface, to a depth referred to as skin depth or penetration depth. Skin depth in Oxygen Free High Conductivity (OFHC) copper at 100°C is 0.1 mm at 600 kHz, and varies inversely with the square root of the frequency. With single-layer windings, current flows only on the surface directly facing the opposing winding – current does not flow on the outer surface. To carry the required current, the conductor must be as wide as possible because of the limited skin depth. The available winding breadth,  $b_w$ , is extremely important.

The design should strive for the minimum number of turns in the windings. Since the number of turns must be an integral number, for low voltage outputs it is often necessary to settle for a single whole secondary turn, even though the calculated ideal would be a fraction of a turn.

Current in the windings always has an AC component, sometimes a DC component as well. If a single layer can not carry the desired AC current because of limited skin depth, making the conductor thicker won't reduce the AC resistance but *will* reduce DC resistance. Multiple thin layers in parallel won't reduce AC resistance – only the layer closest to the opposing winding will carry all of the AC component.

If a winding consists of multiple layers in series, the thickness of each layer must be considerably less than the skin depth. Otherwise, each layer will have *high currents flowing in opposite directions on the two surfaces of each conductor*, and AC losses skyrocket. Thus, multiple series layers cannot be made thicker to improve DC resistance without causing astronomical AC resistance. Reference [2] explains this phenomenon.

### A. Interleaving

Ideally, all windings should have a single layer. A wide winding breadth is essential, as it provides a wide surface area to minimize AC resistance. A single layer can be made much thicker than the skin depth, without adverse effect. This does not further reduce AC resistance, but it does provide lower DC resistance and better heat transfer from the interior of the transformer. A wide winding breadth also reduces leakage inductance by reducing both field intensity (force gradient) and flux density, resulting in a great reduction of energy in the leakage inductance field.

A planar core actually provides *less* winding breadth than a comparable conventional core. However, the windings can be *interleaved*. Interleaving effectively multiplies the winding breadth. Conceptually, a single-layer winding is folded accordion-wise to fit within the available core window. For example, Fig. 16. shows a winding with a single primary layer and single secondary layer folded into six *winding sections*<sup>[2]</sup>. Although there are many layers in the entire structure, each section contains only a single primary and a single secondary layer. *The magnetic field is zero at each section boundary*, so that the field does not build from one section to another. Each layer can then be as thick as desired to reduce DC resistance and improve heat transfer. The interleaved winding of Fig. 16. is functionally equivalent to having a single winding section 6 times wider than the winding breadth available in the core.

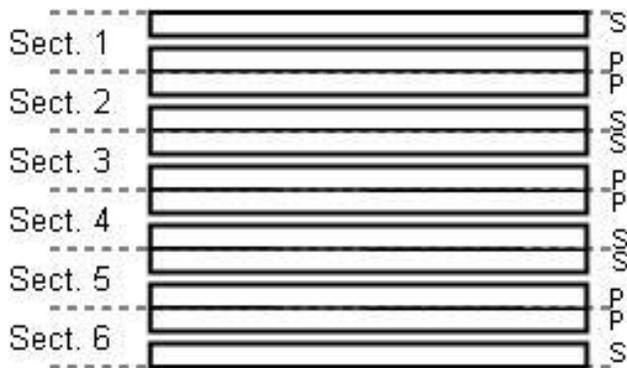


Fig. 16. Interleaved windings.

Interleaving can be carried to a much greater degree in planar devices than with conventional cores, with resulting benefits. This is because the planar winding structure provides much easier access to the windings, simplifying interconnection between layers.

Planar windings can be interconnected in a variety of ways to provide great flexibility. For example, in the winding structure of Fig. 16., there are six primary and six secondary layers – each layer one turn of copper strip or foil, with one layer in each section. The primary layers (single turns) can be connected in various series-parallel combination to provide 1, 2, 3, or 6 turns. The six secondary layers can be connected in the same manner, making possible a wide range of turns ratios. Whenever two adjacent primary or secondary layers are paralleled, they could be replaced by a single turn of twice the thickness. (The section boundary will then be through the center of this single turn.) If all six primary layers/turns are connected in series, then there must be insulation between the adjacent turns.

In another possible variation, individual layers in the interleaved structure could have several turns spiraling out from the center consisting of plated traces on printed wiring boards. If the primary layers in Fig. 16. each consisted of 5 turns, connecting the six layers in series will result in 30 primary turns. If the 6 secondary single-turn layers are then connected in parallel, a 30:1 turns ratio is achieved.

Referring again to Fig. 16., the leakage inductance field between the windings alternates direction in adjacent winding sections. Thus, the leakage flux in Section 1 joins with the flux in Section 2 to form the required complete loops. These flux loops link the leakage inductance energy of Sections 1 and 2 to those windings of Sections 1 and 2 that are enclosed by the flux loops. Likewise, the flux in sections 3 and 4 join together to form complete loops, as does the flux in Sections 5 and 6.

Thus, in a winding structure with an even number of sections, flux loops are completed within adjacent sections. Flux does not extend significantly beyond the extremes of the winding structure. But with a single winding section (or to a lesser degree, with an odd number of winding sections), the flux loops associated with the leakage inductance field between the windings must complete their loops *outside* of the windings, usually to the external portions of the core. But the external paths are long, especially when the windings extend outside the core. Considerable energy then escapes, creating radiated EMI.

The structure of Fig. 16. could have two primary layers on the outside rather than two secondary layers, as shown. But putting the low voltage secondary layers on the outside reduces radiated EMI.

### ***B. Inter-Winding Capacitance***

Interleaving does have one significant adverse effect – increasing the surface area of the windings proportionately increases interwinding capacitance. This has serious implications in off-line applications, by coupling ever-present AC power line noise through to the power supply output. Inter-winding capacitance also injects conducted EMI back onto the power lines by providing a path for switching transients to ground<sup>[3]</sup>. These problems are less significant in applications operating from a low voltage distribution bus which is inherently quieter, and the lower voltages being switched generate less conducted EMI.

Planar transformers have the reputation of having greater inter-winding capacitance than conventional transformers. This is only because interleaving is usually carried to a much greater degree in planar structures than with conventional transformers. If both transformer types were designed with similar core sizes, and with winding breadth/interleaving giving the same AC resistances and leakage inductances, the inter-winding capacitances would be quite similar.

The following factors influence inter-winding capacitance:

- **Dielectric Constant** – Use insulating material with low dielectric constant. Mylar: 3.1, Teflon: 2.1. Both are available in 1 mil (0.025mm) film thickness. The author has no direct knowledge of the insulating qualities of Teflon film. Dielectric Constant has no effect on leakage inductance or AC winding resistance.
- **Insulation thickness** – Thicker insulation reduces capacitance, but proportionately increases leakage inductance. Insulation thickness has no effect on AC resistance
- **Number of turns** – With single layer windings, minimizing turns (by operating the core at max. flux density) reduces the total winding area, hence reduces capacitance. Minimizing turns also minimizes leakage inductance and AC resistance.
- **Winding breadth** – Increased winding breadth, either directly or by interleaving, increases capacitance but decreases leakage inductance and AC resistance of windings.
- **Faraday shields**, properly employed, reduce inter-winding capacitance problems<sup>[3]</sup>, but this adds considerable complexity to the interleaved winding structure, and the increase of inter-winding volume will cause a proportionate increase in leakage inductance.

### ***C. Other Concerns***

Smaller winding breadth makes it difficult to achieve the creepage and clearance distances mandated for safety in off-line applications (Reference 4). With small cores in low power applications, creepage and clearance distances may take up the entire available winding breadth, leaving no room for the winding. One solution to this problem is to use a pre-impregnated sheet bonded to the PWB substrate. This eliminates creepage and clearance between primary and secondary. (a pre-impregnated sheet consists of a mat of reinforcement fibres impregnated with thermoplastic resin, usually partially polymerized to B-stage.)

The low leakage inductance achievable in a well-designed planar transformer can be totally negated by the inductance of circuit wiring inductance external to the transformer. This is especially important with low-voltage outputs. The external circuit paths for high frequency currents must be laid out with extreme care, with the shortest possible paths, wide conductors, and minimal loop areas. An open circuit wiring loop stores energy within the loop which corresponds to inductance. Close this loop area by superimposing conductors as tightly as possible to minimize wiring inductance.

**D. High Frequency AC Current**

In switching power supply applications, transformer current waveforms are rich in harmonic content. While these harmonics (including the high-order harmonics associated with the fast edges) might be of concern regarding EMI, they seldom contribute significantly to winding losses.

To be absolutely accurate when calculating winding losses, it is necessary to calculate the amplitude of each harmonic, square it, and multiply by the AC resistance calculated for each harmonic frequency, then add the losses calculated for the fundamental and each harmonic. This is a tedious process, although it can be facilitated through computer software.

With very narrow pulse widths, the fundamental amplitude is small, and the loss contribution due to harmonics is significant. But this is usually not the worst case. The worst case for AC losses is generally near 50% duty cycle, where the fundamental is maximum, but the harmonics are little changed from their values at narrow pulse widths.

At 50% duty cycle, the 3<sup>rd</sup> harmonic contribution to AC loss is approximately 15% of the fundamental contribution. The higher order harmonics are much less significant – their rms currents *squared* fall off much more rapidly than the AC resistance rises. The total error of ignoring the harmonic contribution might be 20% vs. using the fundamental alone. But if the *total* rms AC current of the square wave (including the fundamental and harmonics) is squared and multiplied by the AC resistance at the fundamental frequency, the calculated loss is significantly greater

than a calculation based on the fundamental alone. Thus, most of the error is erased.

The waveform must be separated into its DC and combined AC components, because the DC resistance can be much lower than the AC resistance at the fundamental frequency. (For a forward converter waveform at 50% duty cycle, the DC component is 0.5 times the peak, and the combined AC component is also 0.5 times the peak current.)

This short-cut method may underestimate winding losses by 5-10%, but it’s easier to throw in a “fudge factor” than to use the rigorous method of loss calculation. Bear in mind that the error is very small compared with the perhaps 50% or more uncertainty of the thermal resistance when calculating temperature rise.

Many of the above points will be illustrated in the design examples which follow.

**XII. PRINTED WIRING BOARD COPPER PLATING**

Make sure the plating vendor uses a process which produces high conductivity – equivalent to OFHC copper. Table I shows the resistance of various thicknesses of copper, expressed in milliOhms per square. Sheet copper of various thicknesses can be scaled from the values given in the table.

Amps/mm Trace Width in Table I is a starting point for estimating current handling capability. Skin depth frequency is the frequency at which the conductor thickness equals the skin depth.

**TABLE I. PWB TRACE RESISTANCE**

Copper Weight (Oz.)	Thickness (mils/mm)	mΩ per Sq. (25°C)	mΩ per Sq. (100°C)	A/mm Trace Width	Skin Depth Freq. (kHz)
1	1.4 / 0.035	0.5	0.6	0.25	5700
2	2.8 / 0.07	0.25	0.3	0.5	1300
3	4.2 / 0.1	0.16	0.2	0.75	625
4	5.6 / 0.14	0.12	0.14	1.0	320

Ohms per square is a convenient way of calculating resistance of PWB traces, or of any conductor in sheet form. For example, if a trace is 15 mm wide and 75 mm long, it has a length of  $75/15 = 5$  squares. If the copper thickness is 0.1 mm, with a resistance of 0.16 m $\Omega$  per square, the trace resistance is  $5 \times 0.16 = 80$  m $\Omega$ . Dimensional units are irrelevant, as long as the same units are used for length and width.<sup>[5]</sup>

By remembering that 0.1 mm copper (3 oz) has 0.16 m $\Omega$  per square, this value can be scaled for any copper thickness.

Minimum 15 mil (0.4 mm) trace width and 15 mil (0.4 mm) spacing are practical values for thick electroplated traces.

### XIII. DESIGN STRATEGY

When a magnetic core is underutilized, it will be larger and more costly than it should be for the intended application. A core is fully utilized when it is designed to operate in its application at maximum flux density, and with the windings operating at maximum current density (not necessarily at the same point of operation). A good strategy for achieving this full utilization goal is to begin by calculating – using Faraday’s Law – the minimum number of turns in the winding that will result in operating the core at maximum flux density (limited by either core loss or saturation).

Unfortunately, the trends toward high frequency operation and lower power supply output voltages makes this goal impossible to achieve in most applications, especially at higher power levels. This is because, as a practical matter, windings must have an integral number of turns, and because, due to skin effect limitations, it is difficult to obtain sufficient conductor surface area to carry the required high frequency currents (other than by interleaving). Very often, the core magnetic cross-section area and winding breadth are mismatched for the intended application – the core area is greater than necessary for a single turn low voltage secondary, yet the winding breadth of that core is too narrow to provide the copper surface necessary to carry the high frequency current, and either a larger core or interleaved windings are required.

When the transformer design for buck-derived topologies results in the core being underutilized, with flux swing only a fraction of what it could be, an option worth considering is to operate the same exact design at a lower frequency. This will increase flux swing, but reduce switching losses, reduce AC winding losses, and reduce circuit losses due to leakage inductance and magnetizing inductance. If the output filter capacitor is electrolytic, with impedance dictated by ESR, it will be no larger at the lower frequency. In buck-derived topologies, the filter inductor becomes larger, but in flyback circuits there appear to be only advantages with operation at a lower frequency where the larger flux swings associated with the integral number of turns approach the limits imposed by core loss or saturation.

Also, a different set of problems become dominant with high frequency operation – especially with low output voltages – casting a different light on topological preferences. Optimum flux density becomes difficult to achieve, and skin effect imposes severe limitations on windings.

### XIV. DESIGN EXAMPLES

Rather than showing the design process for a transformer intended for a specific application, a planar ER25 core is selected as a demonstration vehicle. The capabilities and limitations of this core will be shown in transformer designs operating at 500 kHz transformer frequency. Through these demonstrations, the design approach becomes apparent.

The ER25 Core, shown in shown in Fig. 17. is an industry standard core set available from several manufacturers, in a variety of ferrite core materials. Its external size is 25×15×8 mm (1.0 ×0.6×0.32 in).

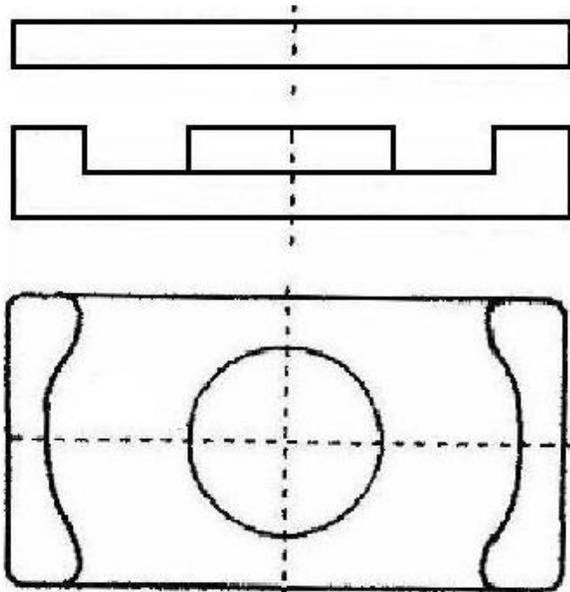


Fig. 17. ER25 core set.

Important characteristics of the ER25 core set are:

Size (Length,Width)	25 × 15 mm
Overall Height	8mm
Magnetic path length: ( $l_e$ )	28.1 mm
Average core area: ( $A_e$ )	$70.4\text{mm}^2 \approx 70 \times 10^{-6} \text{ m}^2$
Core Volume: ( $V_e$ )	$1978\text{mm}^3 \approx 2 \text{ cm}^3$
Winding Breadth (bw)	6.1mm
Winding Height (hw)	3.3 mm
Mean Length of Turn	49 mm

Two high frequency ferrite core materials are considered: Ferroxcube 3F35, and Magnetics “R” material.

#### A. Determine Maximum Flux Swing, $\Delta B$

The first step is to determine the Volt-seconds per turn applied to the ER25 core that will result in operation at maximum flux density (if possible), thus fully utilizing the core. For a filter inductor with small AC current component, maximum flux density is usually limited by saturation, but for a transformer operating at 500kHz, maximum  $\Delta B$  is almost certainly limited by core loss.

*To find maximum  $\Delta B$ :* Refer to the manufacturer’s core loss curves for the core material tentatively selected (see Fig. 18. for Magnetics “R” material). Enter the core loss curve at the desired loss limit – the intersection with the 500 kHz curve shows the corresponding flux density. For a small core, a conservative trial value for core

loss limit is  $150 \text{ mW/cm}^3$  (equivalent to  $150 \text{ kW/m}^3$ ). For the ER25 core, with a core volume of  $2 \text{ cm}^3$ , total core loss at that limit is 0.3 W.

The flux density values given on the core loss curves are peak sinusoidal values. Double these peak values to obtain the corresponding peak-to-peak value,  $\Delta B$ . If the flux density units are given in Gauss, multiply by  $10^{-4}$  to convert to Tesla (calculations use SI units).

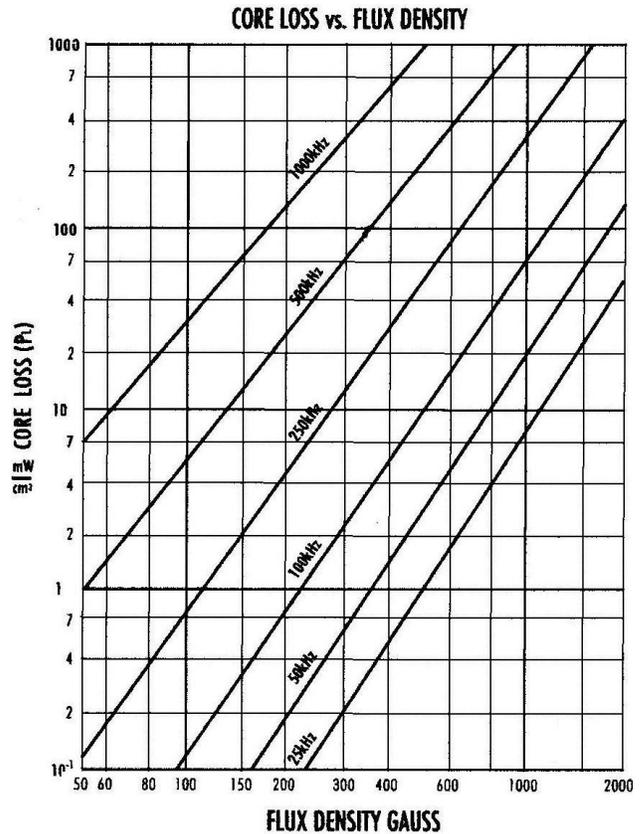


Fig. 18. Core loss curves.

Entering the core loss curves at  $150 \text{ mW/cm}^3$  ( $150 \text{ kW/m}^3$ ), the corresponding  $\Delta B$  limits obtained for the two core materials are:

3F35 material:	$60 \text{ mT(pk)} \times 2 = 0.120\text{T}$
R material:	$40 \text{ mT(pk)} \times 2 = 0.080\text{T}$

Maximum V- $\mu$ -sec per turn – From Faraday’s Law:

$$\frac{d\phi}{dt} = \frac{V}{n}; \quad \frac{V\Delta t}{n} = \Delta\phi = A_e\Delta B$$

For the ER25 core, with  $A_e = 70 \times 10^{-6} \text{ m}^2$ :

3F35 material:

$$0.120T \times 70 \times 10^{-6} = 8.4 \text{ V-}\mu\text{sec per turn}$$

R material:

$$0.080T \times 70 \times 10^{-6} = 5.6 \text{ V-}\mu\text{sec per turn}$$

### B. Forward Converter – Core Utilization

In a forward converter, output voltage,  $V_o$ , equals the secondary V- $\mu$ sec averaged over the switching period. Thus for a single-turn secondary with the ER25 core at 500 kHz (period = 2  $\mu$ s), maximum  $V_o$  equals:

$$\text{3F35 material: } \max V_o = 8.4 \text{ V-}\mu\text{s}/2\mu\text{s} = 4.2 \text{ V}$$

$$\text{R material: } \max V_o = 5.6 \text{ V-}\mu\text{s}/2\mu\text{s} = 2.8 \text{ V}$$

Either material is suitable for a 2.5 V output (2.8V including rectifier drop). 3F35 material is certainly acceptable for 3.3 V (3.6 V) output. If “R” material is used for a 3.3 V output, there are two choices: Either use a single secondary turn, accepting core loss greater than originally planned, or use two turns. With 2 turns, core loss will practically disappear, but the penalty in winding loss is severe – probably much more so than accepting greater core loss with a single turn secondary.

With a 2.5 V output (2.8 V including rectifier), and with a single turn secondary on the ER25 core, the flux swing,  $\Delta B$ , is 80 mT (40mT peak ac), and core loss is 300 mW as calculated above. If 3F35 material is used, core loss is reduced. Working backwards on the core loss curve for 3F35 at 40mT, core loss is 45mW/cm<sup>3</sup>, resulting in a total core loss of only 90 mW.

*Reminder:* The above conclusions apply to the ER25 core operated at 500kHz. With other cores, at other frequencies, the results will be different.

When the secondary voltage is greater than a forward converter transformer can provide with a one-turn secondary (for example, 5V with the ER25 core at 500kHz), consider using a push-pull topology which becomes advantageous at the higher voltage.

### C. Forward Converter Windings

Design of the windings is perhaps the most difficult task in the design of any high frequency transformer, and especially so with planar transformers.

The winding breadth available in planar cores is much less than in conventional cores of similar size. This makes it more difficult to achieve the desired low leakage inductance and AC resistance.

The design attempts to maximize the current output capability of the ER25 core, operating from a 48V telecom distribution bus input, with 2.5 V output,

It is quite possible to complete the design of the secondary winding without any knowledge of the input voltage or the primary/secondary turns ratio. But rather than leave that “loose end,” assume the input from the 48 V telecom bus ranges from 36 V to 72 V. A peak secondary voltage of 6V at the minimum 36V input, with max. duty cycle of 0.47, provides 2.82 V, sufficient for 2.5 V output plus rectifier drop. Thus, a 6:1 turns ratio is required.

How much current can be obtained from the single turn, single layer secondary? The winding breadth of the ER25 core is 6.1 mm. Allowing 0.5 mm each side leaves a trace width of 5 mm. Mean length of the one turn is 49 mm.

The length of the secondary is  $49/5 = \mathbf{10 \text{ squares}}$ .

Skin depth at 500 kHz is 0.112 mm.

Using 4oz. copper (0.14 mm), thickness is slightly greater than the skin depth. Resistance of 4oz. copper is 0.12 m $\Omega$ /square.

Total secondary resistance is  $0.12 \times 10 = \mathbf{1.2 \text{ m}\Omega}$ .

As shown in Fig. 5., the forward converter primary and secondary both have AC and DC current components. Peak secondary current (neglecting ripple) equals DC output current,  $I_o$ . Worst case rms current occurs at low  $V_{IN}$ , when the duty cycle is at or near 0.5. Under this condition, The AC and DC current components each equal one-half of  $I_o$ . Total rms current is  $0.7 I_o$ .

Following the guidelines of Table I, the 5 mm wide 4oz. copper trace should be able to carry 5 Arms, corresponding to  $I_O = 7$  A. AC and DC current components are each 3.5A.

DC loss is  $3.5^2 \times 1.2 \text{ m}\Omega = 14 \text{ mW}$

Since the thickness is very near the skin depth, AC loss is approximately the same: 14 mW.

The primary has exactly the same structure as the secondary, except instead of one turn, there are 6 turns spiraling out from the center across the 5 mm available for the winding. Spacing between turns is 0.4 mm. The primary traces have the same thickness as the secondary, Ampere-turns are the same, current density and skin depth are the same as in the secondary trace. Thus, AC and DC losses in the primary are equal to the secondary losses. Total winding loss, AC and DC, primary and secondary, is approximately 56 mW

The above structure, limited by single layer windings with thickness limited to skin depth, greatly underutilizes available winding space. Two 4-oz copper traces (0.14 mm each) on a 5 mil (0.125 mm) PWB plus 2 layers of 1mil mylar top and bottom sum to a total thickness of 0.5 mm out of the available 3.3 mm winding window height.

#### **D. Double the Windings**

Because the windings are single layer, it is possible to make them thicker without adverse effect. AC resistance, limited by skin depth, remains unchanged but DC resistance is reduced. When multiple traces are connected in parallel, their behavior is identical to a solid layer of equal total thickness.

The secondary can be made up of two 4 oz. traces, paralleled, on each side of a 5 mil PWB, the primary built the same way, separated by 3 mil B-stage pre-impregnated sheet, with two layers of 1 mil mylar top and bottom, for a total thickness of 1.1 mm. DC resistance is cut in half to 0.6 m $\Omega$  and DC loss is halved. Because of limited skin depth, AC loss remains the same. The total winding loss is reduced to 42 mW.

Not only is the loss reduced, but – *this is important* – the additional copper helps remove heat from the interior of the transformer. Load current can be raised to possibly 10 A, depending on the external thermal environment.

The winding window is still greatly underutilized – 1.1 mm is used out of the available 3.3 mm.

#### **E. Interleaving Has Great Benefit**

Interleaving involves duplicating the above single layer structure, arranging the layers either P-S-S-P, or better, S-P-P-S. This structure divides the winding into two winding sections, each section with one layer primary and one layer secondary (each layer has two paralleled traces). Externally, all primary layers are paralleled, and all secondary layers are paralleled. AC resistance is halved to 0.6 m $\Omega$ , DC resistance is 0.3 m $\Omega$ . With 7 A load for each section, 14 A total output current, total winding loss is doubled to 84 mW. If the load current is raised to 20 A, winding loss doubles again to 168 mW,

Total winding height (thickness) is now 2.1 mm out of 3.3 mm available.

*An additional important benefit of this first level of interleaving is that leakage inductance is halved, (although inter-winding capacitance is doubled).*

#### **F. Copper Strip Secondary**

With the secondary layers on top and bottom of the winding structure, further great improvement can be achieved by making them thicker, utilizing the additional space available in the window. For example, each of the two secondary layers could be made from 1 mm copper stampings. AC resistance, defined by skin depth, would not be reduced, but DC secondary loss is almost eliminated. Total winding loss at 20 A output current is 145 mW. More importantly, the thick copper secondaries are very effective in removing heat from the transformer interior. It is not unreasonable to extend the output current capability to 25 A, given a suitable thermal environment. The winding structure is shown in Fig. 19.

Winding losses calculated above are maximum, occurring at minimum  $V_{IN}$ , and 0.47 maximum duty cycle. Losses at nominal  $V_{IN}$  are less.

Layer	Thickness, mm
Mylar, 2x1mil	0.05
Copper	1.0
B-stage Adhesive	.076
Plated copper, 4oz.	0.14
Printed Wiring Board	0.125
Plated copper, 4oz.	0.14
Mylar, 1 mil	.025
Plated copper, 4oz.	0.14
Printed Wiring Board	0.125
Plated copper, 4oz.	0.14
B-stage Adhesive	.076
Copper	1.0
Mylar, 2x1mil	0.05
<b>Total Winding Height</b>	<b>3.1 mm</b>

Fig. 19. Transformer winding structure.

Core loss depends only on Volt- $\mu$ sec per turn and frequency. For the buck derived topologies,  $V\text{-}\mu$ s are constant in steady-state operation – independent of input voltage and load current.

If greater output current capability is required, interleaving the windings can be carried further, to include one primary layer and one secondary layer in each section. Each layer can then be made considerably thicker than the skin depth without penalty, reducing DC loss and facilitating heat removal. Primary and secondary layers can be connected in series for higher voltage or in parallel for higher current, as desired.

For example, the interleaved structure of Fig. 16. could be employed in the previous design example. The six single-turn secondary layers could be externally paralleled, and the six single turn primary layers externally series connected, resulting in a 6:1 turns ratio. Leakage inductance and AC resistance are dramatically reduced. The 6 winding sections shown in Fig. 16. produce the same result as having a single section with six times the winding breadth as the core actually provides.

But the penalty for extensive interleaving is a proportional increase of inter-winding capacitance, which couples input noise and primary-side switching transients to the output. If the need for low inter-winding capacitance outweighs the need for reduced AC resistance and leakage inductance, it is necessary to use several spiraled turns in each layer so as to achieve the required number of turns with fewer interleaved layers.

**Connecting to the center of the spiral:** In the previous design example, how is the connection made from the external circuit to the inner end of the 6-turn spiral primary winding? The simple answer is to bring additional conductors from the inner ends of the spiral winding to the outside. But this requires more layers of copper and insulation, making the total winding thicker than desired.

Another, more elegant solution is as follows: Instead of a 6-turn spiral (4 traces, in 2 layers, 2 winding sections, all paralleled), make two identical layers (each with 2 traces) of 3 turns each, each turn twice the width of the original 6-turn spiral. Locate the inner ends of these 3-turn spirals centrally, so that when one of the two layers is flipped over, its spiral inner end is directly over the inner end of the other winding layer. Fig. 19. shows these two layers side-by-side (although they will actually be superimposed). All of the traces at the center of the spirals can then be interconnected through vias.

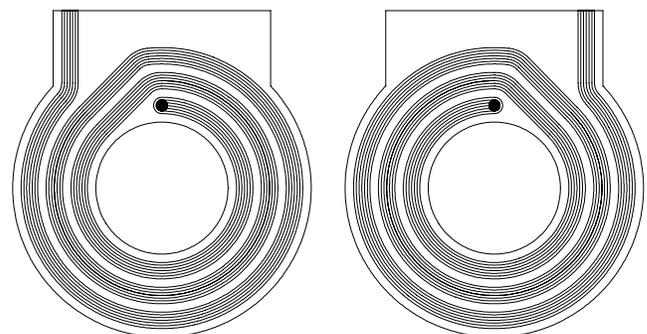


Fig. 20. Series-connected 3-turn spiral windings.

The result is a six-turn series connected primary, with 3 turns in each of two layers. The two series-connected layers are in two different winding sections (which is essential for proper functioning). The outside ends of the series connected primary layers are at different corners, facilitating connection to the external circuit. No additional connection to the center of the spiral windings is required.

### G. Mounting the Transformer

Instead of mounting the planar transformer upon, or against, the external circuit board, it is best to mount it through a cutout in the board. This facilitates interconnection between the windings and the external board, and makes much more core surface area available for cooling. Ultimately, the current handling capability of any transformer design depends upon how effectively its thermal environment removes heat, to keep temperature rise within acceptable limits.

The ER25 core, using 3F35 material has the  $V\text{-}\mu\text{s}$  capability to operate at 500 kHz with a 3.3 V output. The 2.5 V design as implemented above (perhaps optimized by using five primary turns instead of six) can deliver the same current at 3.3 V. Thus, greater power output capability is achieved by more fully utilizing the core.

The single-ended forward converter topology is attractive because of its simplicity. However, using the ER25 core at 500 kHz with a single turn secondary, the forward converter is limited to an output voltage below 3.9 V. Above that voltage, a core with greater area,  $A_e$ , is necessary to obtain the required  $V\text{-}\mu\text{s}$  capability, or requires multiple series-connected secondary layers in an appropriately interleaved winding structure.

## XV. HALF-BRIDGE TOPOLOGY

The push-pull buck-derived topologies such as the half-bridge, full-bridge, and push-pull center-tap have one-half the secondary  $V\text{-}\mu\text{s}$  as the forward converter, for the same output voltage. For the same core loss, the ER25 core with single turn secondary could provide up to 8 V output in a push-pull circuit. However, push-pull circuits have twice the high frequency AC current in the windings as the forward converter for the same DC output current. At low frequencies this is not a problem – the push-pull circuits would have half as many secondary turns of thicker wire, one-fourth the resistance, and could therefore handle twice the current. But when the secondary is at the minimum of one turn, the push-pull circuits, using the same core, are limited to the same AC secondary current as the forward converter, resulting in half the DC output current capability. So, at high frequency, the push-pull topologies are disadvantaged, and their greater power handling capability is not achievable. This problem can be resolved through interleaving.

If a push-pull topology is used to obtain higher output voltage capability than the forward converter with a single turn, single layer secondary, the half-bridge is the best choice. Center-tap windings significantly reduce current-handling capability and should be avoided. This is because the center-tap configuration puts additional layers within the winding structure. At a time when one side of the center-tap winding is conducting, the other side has zero external current. Nevertheless, the non-conducting portion acts as a *passive layer* within the winding structure. Unless it is very thin, it will have equal and opposite currents flowing on its two surfaces, resulting in significant loss, even when not conducting external current<sup>[2]</sup>. With an interleaved structure, center-tap windings must therefore be limited in thickness to the skin depth, or each side of the center-tap winding must be in its own winding section, making for a bulkier winding.

While the half-bridge avoids using a center-tap primary, avoiding a center-tap secondary requires bridge rectification, which benefits transformer design, but hurts circuit efficiency.

## XVI. CURRENT DOUBLER SECONDARY

The topological combination which provides the best transformer utilization is the bridge or half-bridge primary with current doubler secondary<sup>[6]</sup>. As shown in Fig. 21., the current doubler requires two (smaller) filter inductors, but has a single-winding secondary and uses only two rectifiers, thus avoiding the rectifier losses of full-bridge rectification. For the same output voltage, secondary  $V\text{-}\mu\text{s}$  per turn are the same as the forward converter and twice that of the center-tap secondary. Like the forward converter, rms AC secondary current equals one-half of output current,  $I_O$ , but *unlike* the forward converter, DC secondary and primary currents are zero

The forward converter transformer design accomplished earlier in this paper could be used without change in the same application but with full-bridge primary and current doubler secondary. It would have the same rms AC primary and secondary currents as the forward converter, but zero DC primary and secondary current components. Without DC losses, output current could be pushed to greater levels.

Current doubler output ripple frequency is twice the transformer frequency – twice the output frequency of the forward converter. This enables smaller filter inductor values and enhances the possibility of replacing an electrolytic output filter capacitor with a smaller and less expensive ceramic capacitor.

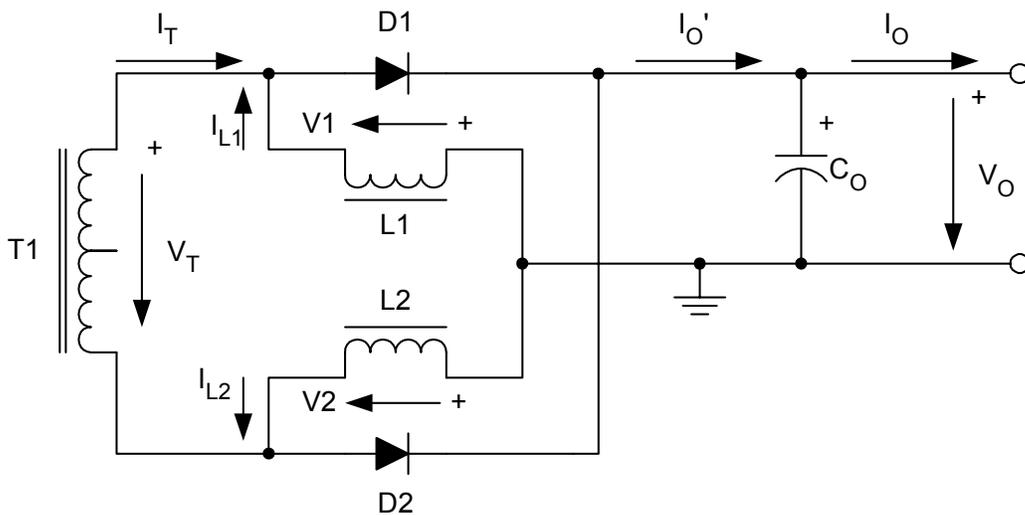


Fig. 21. Current doubler.

## XVII. SUMMARY

Planar transformers have many advantages over their conventional counterparts – lower profile, better heat removal, fewer turns, ease of interleaving, smaller leakage inductance, lower winding resistance, and good reproducibility.

Understanding magnetic field principles and behavior is essential for achieving optimized transformer design.

The planar structure is advantageous in a “true” transformer because the AC magnetic field “pulls” the AC current to the broad conductor surfaces.

The planar structure is disadvantageous in a discontinuous flyback transformer because the AC field through the gap pulls AC current to the narrow edges of the conductors facing the gap.

Use a single layer in all windings. Single layers permit windings to be much thicker than skin depth, minimizing DC resistance and facilitating heat removal.

If more than one layer is required, interleave the windings, which effectively results in single layers in each winding section. This greatly reduces AC resistance and leakage inductance, but increases inter-winding capacitance.

Consider using a full/half bridge primary with current doubler secondary topology.

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## Appendix

### Magnetic Field Characteristics

A magnetic field is a form of stored energy. It consists of two components – Magnetic Force,  $F$ , and Magnetic Flux,  $\Phi$ .

- In a transformer or inductor, magnetic force is derived from electrical current. In the SI system of units, the unit of magnetic force,  $F$ , equals the current producing the field, in Amperes (Ampere's Law):

$$F = I \quad (\text{Amperes})$$

- Magnetic force can be described as a series of equipotential surfaces. These finite surfaces are bounded by, and terminate upon, the current which produces the field.
- The magnetic force gradient,  $H$  (usually called *Field Intensity*), at any point is defined by the distance between the equipotential surfaces:

$$H = dF/d\ell \quad (\text{Amps/meter})$$

- The total magnetic force integrated along *any* closed path is equal to the total current enclosed by that path:

$$F = \int H d\ell = I \quad (\text{Amperes})$$

- Magnetic flux,  $\Phi$ , can be described as a bundle of lines. Flux lines always form closed loops – no flux line has a beginning or an end. Flux lines intersect the magnetic force equipotential surfaces, and are always normal (at right angles to) these surfaces at the points of intersection.

- Flux density,  $B$ , at any point, is defined as the amount of flux per unit cross-section area at that point:

$$B = d\Phi/dA \quad (\text{Webers/m}^2)$$

- Flux density at any point equals field intensity,  $H$ , multiplied by the magnetic permeability of the material containing the field at that point.

$$B = \mu H \quad (\text{Webers/m}^2)$$

- In the SI system of units, permeability,  $\mu$ , has two components –  $\mu_0$ , absolute permeability, and  $\mu_r$ , relative permeability

$$\mu_0 = \mu_0 \mu_r$$

- Permeability can never be zero. Absolute permeability,  $\mu_0$ , is the permeability of free space or any non-magnetic material (air, insulating materials, copper), and is essentially the minimum permeability value. In the SI system:

$$\mu_0 = 4\pi \cdot 10^{-7}$$

- Relative permeability for any non-magnetic material equals 1. For magnetic materials,  $\mu_r$  establishes their greater permeability value.
- The energy density,  $W/v$ , of the magnetic field at any point equals the area between the B-H characteristic and the B axis, integrated from zero to the operating point

$$W/v = \int H dB \quad (\text{Joules/m}^3)$$

- When permeability is constant, as in any non-magnetic material, B and H are linearly related, and:

$$W/v = \frac{1}{2} BH \quad (\text{Joules/m}^3)$$

- Total energy stored in any portion of a magnetic field equals the energy density integrated over the volume of the field portion:

$$W = \int H \partial B \partial v \quad (\text{Joules/m}^3)$$

- When the field is in a region of constant permeability and constant field intensity, such as in an air gap or the inter-winding region in a transformer or inductor, the above equation simplifies to:

$$W = \frac{1}{2} BHv \quad (\text{Joules/m}^3)$$