CHAPTER 12

Forward and Flyback Converters: Step-by-Step Design and Comparison

Introduction

This is a follow-up chapter to everything we have learned in previous chapters on AC-DC design in particular. It brings it all together in a top-down numerical example. Certainly a level of expert knowledge obtained from previous chapters will be required, but we are trying to keep it very accessible, so we may repeat previous learnings.

In a Flyback topology, the selection of the transformer core is fairly straightforward. The Flyback transformer has a dual function: It not only provides a step-up or step-down ratio based on the Primary-to-Secondary turns ratio, but it also serves as a medium for energy storage. The Flyback is a derivative of the Buck-boost and shares its unique property that not just part but *all* of the energy that is delivered to the output must have previously been stored (as magnetic energy) within the core. This is consistent with the fact that the secondary winding conducts only when the primary winding stops, and vice versa. We can intuitively visualize this as the windings being *out of phase*. So we have an endless sequence of energy stored-and-released followed by stored-and-released, and so on. The core-selection criterion is thus very simply as follows: The core must basically be capable of storing each packet of energy (per cycle) passing through it. That packet is equal to $P_{IN}/f = \Delta\varepsilon \approx \varepsilon_{PEAK}/1.8 = (L \times I_{PEAK}^2)/3.6$, in terms of joules. Here f is the switching frequency and ε is energy (see Fig. 5.6 of *Switching Power Supplies A-Z*, 2d ed, for a derivation of the preceding). Equivalently, we can just state that the peak current I_{PEAK} should not cause *core saturation*, though that approach gives us no intuitive understanding of the fact that if we double the switching frequency, the energy packets get reduced in half, and so in effect the same core, designed properly, can handle twice the input-output energy. But that is indeed always true whenever we use an inductor or transformer as an energy-storage medium in switching power conversion.

Coming to a Forward converter, at least two things are very different right off the bat.

1. Not all the energy reaching the output necessarily needs to get stored in a magnetic energy storage medium (core) along the way. Keep in mind that the Forward converter is based on the Buck topology. We realize from p. 208 of *Switching Power Supplies A-Z*, 2d ed, that only $1 - D$ times the total energy gets cycled through the core in a Buck topology. So, for a given P_O, and a given switching frequency, the Buck or Forward core will be roughly half the size of a Buck-boost or Flyback core, handling the same power (assuming $D \approx 1 - D \approx 0.5$).
2. Further, in a Forward converter, the energy storage function does not reside in the transformer. The storage requirement, however limited, is fulfilled entirely by the secondary-side choke, not the transformer. So we can well ask: What does the transformer do in a Forward converter anyway? It only provides *transformer action*, i.e., voltage step-up corresponding to current step-down or voltage step-down corresponding to current step-up function, based on the turns ratio—which is, in a way, half the function of a Flyback transformer. Once it provides that step-up or step-down ratio, there is an additional step-down function provided by simply running the secondary-side choke in a chopped-voltage fashion, as in any regular (nonisolated) Buck. That is why we always consider the output rail of a Forward

converter as having been derived from the input rail, with two successive step-down factors applied, as shown

$$V_O = (D \times V_{\text{IN}}) \quad \times \quad \frac{N_S}{N_P}$$
$$\Uparrow \qquad\qquad \Uparrow$$
$$\text{Buck} \qquad \text{Transformer action}$$

The perceptive will notice that the Forward converter's transformer action could be such that we use the transformer turns ratio to give an intermediate step-up instead of a step-down function, and then follow it up with a step-down function accruing from the inherent Buck stage based around the secondary-side choke. That could in effect give us another type of (overall) Buck-boost converter—but not based on the classic inductor-based Buck-boost anymore. And that is what we, in effect, usually do in the LLC resonant topology (see Chap. 19).

The secondary-side choke selection criterion is straightforward too: It is sized so that it does not saturate with the peak current passing through it (typically 20 percent more than the load current). We see that it is the same underlying criterion as in a Flyback, Buck, Buck-boost, and even a Boost. So this leaves us with the basic question: How do we pick the Forward converter *transformer*? What does its size depend on? What are its selection criteria?

There are two major factors affecting the Forward converter transformer selection. First, we need to understand that the primary and secondary windings conduct *at the same time*. So they are intuitively "in phase." The observed *transformer action*, that is, the simple turns-ratio–based current flow of the secondary winding, is in fact just a direct result of induced electromotive force (EMF, i.e., voltage) based on Faraday's and Lenz's laws. The induced EMF in the secondary winding in response to the changing flux caused by the changing current in the primary winding tries to oppose the change of flux, and since both windings can conduct current at the same time in a Forward converter, the two voltages (applied and induced) lead to simultaneous currents in the windings, which create equal and opposite flux contributions in the core, cancelling each other out. Yes, completely so! In effect, the *core* of the Forward converter's transformer does not "see" *any* of the flux associated with the transfer of power across its isolation barrier. Note that this flux-cancellation "magic" was physically impossible in a Flyback, simply because, though there was induced EMF in the Secondary during the ON-time, the output diode was so pointed that it blocked any current flow arising from this induced voltage—so there was no possibility of having two equal and opposite flux contributions occurring (at the same time).

This leads to the big question: If the *core* of the Forward converter's transformer does not see any of the flux related to the ongoing energy transfer through the transformer, can we transfer limitless energy through the transformer? The answer is no, because the DC resistance of copper gets in the way. This creates a *physical limitation* based on the available window area W_a of the core. We just cannot stack endless copper windings in a restricted space to support any power throughput. Certainly not if we intend to keep to certain thermal limits; because though the core may be totally "unaware" of the actual currents in the windings (because of flux cancellation), the windings themselves do see I^2R (ohmic) losses. So eventually, for thermal reasons, we have to keep to within a certain acceptable *current density*. This in effect restricts the amount of power we can transfer through a Forward converter transformer. We intuitively expect that if we double the available window area W_a, we would be able to double the currents (and the power throughput) too, for a given (acceptable) current density. In other words, we expect roughly (intuitively)

$$P_O \propto W_a$$

Truth does in fact support intuition in this case. But there is another key factor too: A transformer needs a certain excitation (magnetization) current to function to be able to provide transformer action in the first place. So there is a certain relationship to the core itself, its *ferrite-related* dimensions, not just the window area (air dimensions) that it provides. A key parameter that characterizes this aspect of the core is the area of its center limb, or A_e (often just called A in this chapter). Finally we expect the power to be related to both factors: the air-related component W_a and the ferrite-related component A_e:

$$P_O \propto W_a \times A_e$$

The product $W_a \times A_e$ is generically called *AP*, or the area product of the core. See Fig. 12.1.

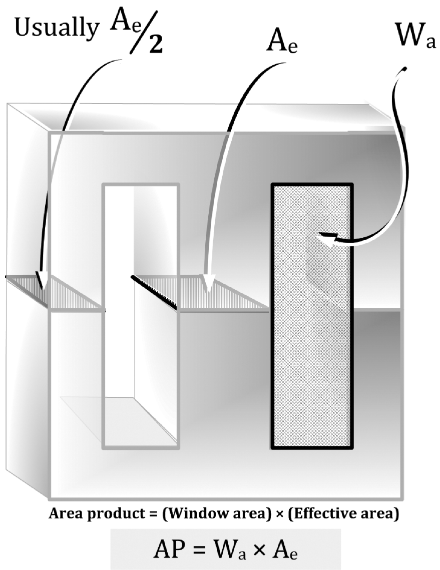


Figure 12.1 Basic definition of area product.

As indicated, we intuitively expect that doubling the frequency f will double the power too. So we expect

$$P_O \propto \mathrm{AP} \times f$$

Or better still, since in the worst case (losses after the transformer) the transformer is responsible for the entire *incoming* power, it makes more intuitive sense to write

$$P_{\mathrm{IN}} \propto \mathrm{AP} \times f$$

Finer Classes of Window Area and Area Product (Some New Terminology)

As we can see from Figs. 12.2 and 12.3, we can actually break up the window area into several windows (with associated area products). We should try to distinguish between them for the subsequent analysis, since typically this becomes a source of major confusion in literature, with innumerable equations and fudge factors (generically called K_x usually) being used

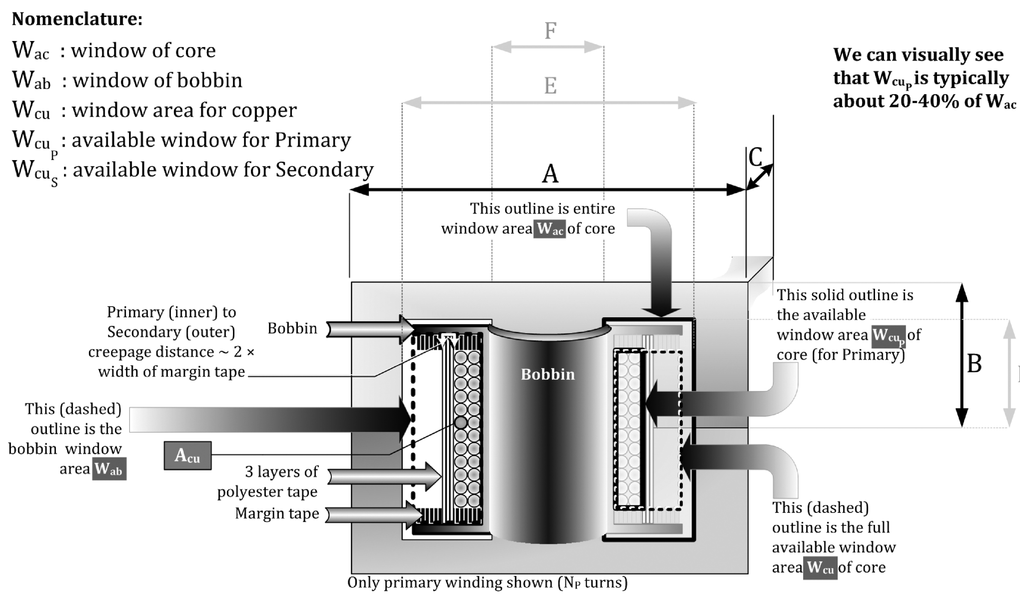


Figure 12.2 Finer divisions of window area and area product.

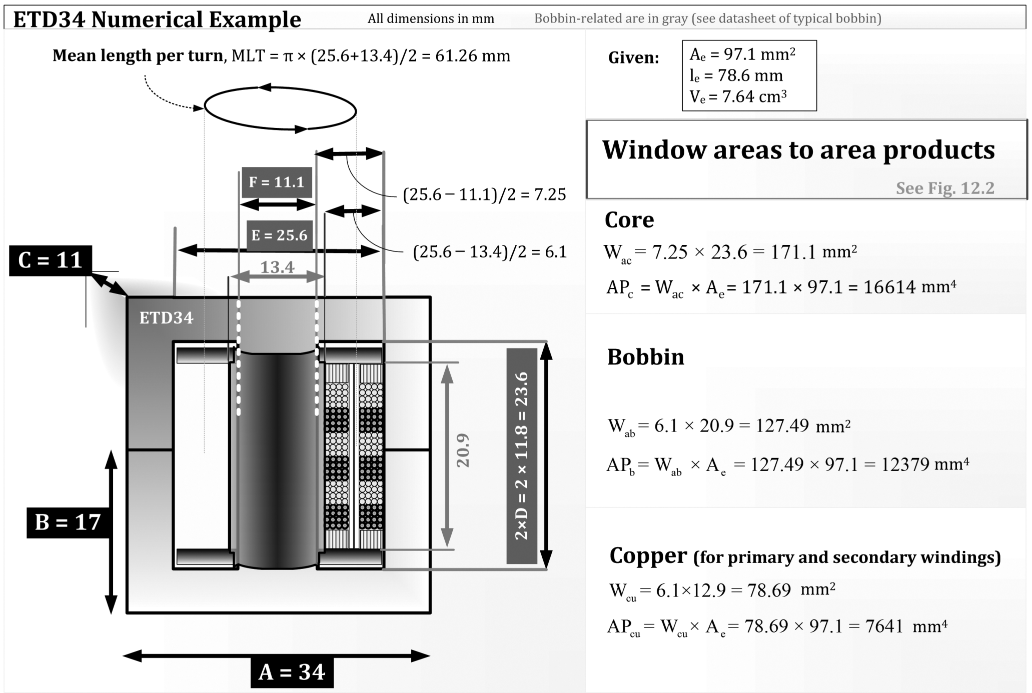


Figure 12.3 Numerical example showing the nomenclature of popular dimensions and also various window areas and area products.

apparently to fit equations somehow to empirical data, rather than deriving equations from first principles and then seeing how they match the data. So we create some descriptors here.

W_{ac}. This is the core window area. Multiplied by A_e, we get AP_c.

W_{ab}. This is the bobbin window area. Multiplied by A_e, we get AP_b.

W_{cu}. This is the window available to wind copper in (both primary and secondary windings). Multiplied by A_e, we get AP_{cu}.

Note *In a safety-approved transformer for AC-DC applications, we typically need 8-mm creepage between primary and secondary windings (see Fig. 12.2 and Fig. 17.1), so a 4-mm margin tape is often used (but sometimes only 2.5 to 3 mm wide nowadays). For telecommunication applications, where only 1500 Vac isolation is required, a 2-mm margin tape will suffice and provide 4 mm of creepage. The bobbin, insulation, etc., significantly lower the available area for copper windings—to about 0.5 × (or half) the core window area* W_{ac}.

W_{cu_P}. This is the window available for the primary winding. Multiplying it by A_e, we get AP_{cu_P}. For a safety-approved AC-DC transformer, for example, this area may be only 0.25 times W_{ac} (typically assuming W_{cu} is split equally between the primary and secondary windings).

W_{cu_S}. This is the window available for the secondary winding. Multiplying it by A_e, we get AP_{cu_S}.

Power and Area Product Relation

We remember that since the voltage across the inductor during the ON-time, V_{ON}, equals the input rail V_{IN} in *almost* all topologies (though not in the half-bridge, for example), from the original form of the voltage-dependent (Faraday) equation

$$\Delta B = \frac{V_{IN} \times t_{ON}}{N_P \times A} \quad \text{T}$$

Here A is the effective area of the core (same as A_e), expressed in square meters. (To remember try this: "volt-seconds equals NAB"). Note that

$$N_P \times A_{cu} = 0.785 \times W_{cu_P}$$

This is because a round wire of cross-sectional area A_{cu} occupies only 78.5 percent (i.e., $\pi^2/4$) of the space (square of area D^2) that it physically occupies within the layer (see Fig. 10.3). Here W_{cu_p} is the (rectangular) physical window area available to wind copper in, but is reserved only for the primary turns. We are typically assuming that the available copper space W_{cu} is split equally between primary and secondary windings. That is a valid assumption mostly.

Solving for N_P, the number of primary turns is

$$N_P = \frac{0.785 \times W_{cu_P}}{A_{cu}}$$

Using this in the voltage-dependent equation, we get

$$\Delta B = \frac{V_{IN} \times t_{ON} \times A_{cu}}{0.785 \times W_{cu_P} \times A_e} \quad \text{T}$$

Performing some manipulations

$$\Delta B = \frac{V_{IN} \times t_{ON} \times A_{cu}}{0.785 \times W_{cu_p} \times A_e} = \frac{V_{IN} \times (I_{IN}/I_{IN}) \times (D/f) \times A_{cu}}{0.785 \times W_{cu_p} \times A_e}$$
$$= \frac{P_{IN} \times D \times A_{cu}}{I_{IN} \times 0.785 \times W_{cu_p} \times A_e \times f} = \frac{P_{IN} \times D \times A_{cu}}{(I_{SW} \times D) \times 0.785 \times W_{cu_p} \times A_e \times f}$$
$$= \frac{P_{IN}}{(I_{SW}/A_{cu}) \times 0.785 \times W_{cu_p} \times A_e \times f} = \frac{P_{IN}}{J_{A/m^2} \times 0.785 \times AP_{cu_p} \times f}$$

where $J_{A/m2}$ is the current density expressed in A/m^2 and AP_{cu_P} is the *area product* for the copper allocated to the primary windings ($AP_{cu_P} = A_e \times W_{cu_P}$). Note that I_{SW} *here is the center of ramp (COR) of the switch current (its average value during the ON-time). The current density is, therefore inherently based on that, not on the RMS current*, as is often erroneously interpreted. Let us now convert the preceding into CGS units for convenience (writing units explicitly in the subscripts to avoid confusion). We get

$$\Delta B_G = \frac{P_{IN}}{J_{A/cm^2} \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

where AP_{cu_P} is expressed in square centimeters now. Finally, converting the current density into cmil/A (see Table 10.1) by using

$$J_{cmil/A} = \frac{197{,}353}{J_{A/cm^2}}$$

we get

$$\Delta B_G = \frac{P_{IN} \times J_{cmil/A}}{197{,}353 \times 0.785 \times f_{Hz} \times AP_{cu_{P\_cm^4}}} \times 10^8 \quad \text{G}$$

Solving for the Primary copper area product

$$AP_{cu_{P\_cm^4}} = \frac{645.49 \times P_{IN} \times J_{cmil/A}}{f_{Hz} \times \Delta B_G} \quad \text{cm}^4$$

Let us do some numerical substitutions here. Assuming a typical target current density of 600 cmil/A (based on COR current value as previously explained), and a typical allowed ΔB equal to 1500 G (to keep core losses down and to avoid saturation), we get the following core selection criterion:

$$AP_{cu_{P\_cm^4}} = 258.2 \times \frac{P_{IN}}{f_{Hz}} \quad \text{cm}^4 \qquad \text{(for 600 cmil/A, based on center of current ramp)}$$

Keep in mind that so far this is an exact relationship. It is based on the window area available for the primary winding, because, along with the target current density in mind (600 cmil/A), this determines the ampere-turns and thus the flux.

In p. 153 of *Switching Power Supplies A-Z*, 2d ed, we derived the following relationship in a similar manner to what we have done here:

$$P_{IN} = \frac{AP_{cm^4} \times f_{Hz}}{675.6}$$

Equivalently

$$\mathrm{AP}_{\mathrm{cm}^4} = 675.6 \times \frac{P_{\mathrm{IN}}}{f_{\mathrm{Hz}}}$$

This too was based on a COR current density of 600 cmil/A. The real difference with the equation we have just derived is that the area product in the *A-Z* book used the entire core area. In other words we had derived this:

$$\mathrm{AP}_{c_{\mathrm{cm}^4}} = 675.6 \times \frac{P_{\mathrm{IN}}}{f_{\mathrm{Hz}}}$$

Compared to what we just derived (based on estimated area reserved for the primary winding)

$$\mathrm{AP}_{\mathrm{cu}_{P\_\mathrm{cm}^4}} = 258.2 \times \frac{P_{\mathrm{IN}}}{f_{\mathrm{Hz}}}$$

In effect we had assumed in the *A-Z* book that $\mathrm{AP}_{\mathrm{cu}_P}/\mathrm{AP}_c = 258.2/675.6 = 0.38$. (*Note*: The reason it seems to be set to 0.3 in the *A-Z* book is this: 0.3/0.785 = 0.38! Think about it. There is no contradiction. The factor 0.785 was not factored into the current density.) But, in the *A-Z book*, as in most literature, the utilization factor *K* was just a fudge factor, applied to make equations fit data (with some physical reasoning to satisfy the critics). But in our ongoing analysis, we are trying to avoid all inexplicable fudge factors. So we should assume the equation we have just come up with is accurate.

Keep in mind that though the maximum flux swing of 1500 G is still a very fair assumption to still make in most types of practical Forward converters (to limit core losses and avoid saturation during transients), the current density of 600 cmil/A (COR value) needs further examination. And until we do that, let us stick to the more general equation connecting area product and power (making no assumptions yet).

$$\mathrm{AP}_{\mathrm{cu}_{P\_\mathrm{cm}^4}} = \frac{645.49 \times P_{\mathrm{IN}} \times J_{\mathrm{cmil/A}}}{f_{\mathrm{Hz}} \times \Delta B_{\mathrm{G}}} \quad \mathrm{cm}^4 \quad \text{(Maniktala, most general)}$$

In terms of A/cm², this is

$$\mathrm{AP}_{\mathrm{cu}_{P\_\mathrm{cm}^4}} = \frac{645.49 \times P_{\mathrm{IN}} \times 197{,}353}{f_{\mathrm{Hz}} \times \Delta B_{\mathrm{G}} \times J_{\mathrm{A/cm}^2}}$$

or

$$\mathrm{AP}_{\mathrm{cu}_{P\_\mathrm{cm}^4}} = \frac{12.74 \times P_{\mathrm{IN}}}{f_{\mathrm{kHz}} \times \Delta B_{\mathrm{T}} \times J_{\mathrm{A/cm}^2}} \qquad \text{(Maniktala, most general)}$$

Keep in mind that *J* here is based on the COR value.

Current Density and Conversions Based on *D*

At the very start of the preceding derivation, when we set $I_{\mathrm{IN}} = I_{\mathrm{SW}} \times D$, in effect the current density was a *COR* current density, not an RMS value. That is how we *eliminated D* from the equation. However, heating does not depend directly on the COR value, but on its RMS. So, in effect, looking at it the other way, our area product equation actually implicitly *depends on D* through the COR current density value we picked. If we know *D*, we can convert the COR-based current density to an equivalent RMS current density value.

The 600-cmil/A value we used to plug in numerically into the equation should perhaps be written out more clearly as 600 cmil/A_{COR}, where A_{COR} is the center-of-ramp value of the current in amperes. We ask: What is 600 cmil/A in terms of RMS current? As indicated, that actually depends on the duty cycle. Assuming a ballpark nominal figure of $D = 0.3$ for a Forward converter, a current pulse of height 1 A leads to an RMS of $1\,\mathrm{A} \times \sqrt{D} = 1\,\mathrm{A} \times \sqrt{0.3} = 0.548$ A. In other words, 600 cmil/A_{COR} means that 600 cmil is being allocated for 0.548 $\mathrm{A_{RMS}}$. In other words, this is equivalent to allocating 600/0.548 = 1095 cmil per $\mathrm{A_{RMS}}$. So we get the following conversions:

$$\frac{600\ \mathrm{cmil}}{\mathrm{A_{COR}}} \equiv \frac{600}{0.548} = \frac{1095\ \mathrm{cmil}}{\mathrm{A_{RMS}}}$$

In other words, 600 cmil/A_{COR} can be expressed as 1095 cmil/A_{RMS}, or

$$\frac{197{,}353}{600} = \frac{330 A_{COR}}{cm^2} \quad \text{(in terms of COR current, see Table 10.1)}$$

or

$$\frac{197{,}353}{1095} = \frac{180\ A_{RMS}}{cm^2} \quad \text{(in terms of RMS current, for } D = 0.3\text{)}$$

Note that we were in effect asking for a current density of 180 A/cm^2, which is rather lower (more conservative) than usually accepted. But let us discuss this further below.

Optimum Current Density

What really is a good current density to target in an application? Is it 600 cmil/A_{COR} (i.e., 180 A_{RMS}/cm^2 for $D = 0.3$) or something else? Actually, 600 mil/A_{COR} is a tad too conservative as we too will agree here. But, in general, this is a topic of great debate, much confusion, and widely dissimilar recommendations in the industry. We need to sort it out.

As a good indication of the industrywide dissonance on this subject, see the 40-W Forward converter design from an engineer at Texas Instruments at www.ti.com/lit/ml/slup120/slup120.pdf. He writes,

> The transformer design uses the Area Product Method that is described in [3]. This produced a design that was found to be core loss limited, as would be expected at 200 kHz. The actual core selected is a Siemens-Matsushita EFD 30/15/9 made of N87 material. The area-product of the selected core is about 2.5 times more area-product than the method in [3] recommended. We selected the additional margin with the intention of allowing additional losses due to proximity effects in a multi-layer foil winding that is required for carrying the large secondary currents.

In this extract, the engineer refers to reference [3], which is: Lloyd H. Dixon's "Power Transformer Design for Switching Power Supplies," Rev. 7/86, *SEM-700 Power Supply Design Seminar Manual*, Unitrode Corporation, 1990, section M5.

This means that *Unitrode [now Texas Instrument (TI)] has a recommendation on core size of Forward converters that was almost 250 percent off the mark, as reported by another TI engineer who actually tried to follow his own company's design note to design a practical converter.*

It therefore seems it is a good idea to stay conservative here, as no one in the commercial arena will appreciate or reward a thermal (or EMI) issue holding up safety approvals and production at the very last moment.

Let us start with the basics: It has been stated and seemingly widely accepted that for most E-core–type Flyback (not Forward) transformers, a current density of 400 cmil/A_{RMS} (equivalent to 197,353/400 ≈ 500 A_{RMS}/cm^2) is acceptable. This seems to have at least served engineers making evaluation boards well. But is it really acceptable in trying to achieve a maximum 55°C rise (internal hot-spot temperature), so as to qualify as a commercial safety-approved Class A Forward converter transformer (maximum of 105°C)?

The problem is that a current density of 500 A_{RMS}/cm^2 may work for low-frequency sine waveforms, as used by most core vendors, but when it comes to Forward converters in particular, because of the skin and proximity effects, as best described by Dowell historically, the ratio F_R (AC resistance divided by DC resistance) is much higher than unity. Note that Dowell used high-frequency waves for a change, but still assumed sinusoidal waves. After that, a lot of Unitrode application notes invoked the original form of Dowell's equations, with sine waves, and arrived at "achievable" F_R values slightly greater than 1, with proper high-frequency winding techniques, and so on. However, in modern days, when we include the high-frequency harmonics of the typical *square waveforms* of switching power conversion, the best achievable AC resistance ratio F_R is not close to 1, but about 2. In other words, mentally we can think of this as windings made with a new metal that has double the resistivity of copper. Now, to arrive at the same acceptable value of heating and temperature rise as regular (low-frequency) *copper transformers*, a good target in a Forward converter would be to allocate twice the area (i.e., target half the current density expressed in A/cm^2). That means we really want to target 800 cmil/A_{RMS} for a Forward converter, which would be roughly comparable in temperature to 400 cmil/A_{RMS} for a Flyback. So, assuming a Forward converter with $D = 0.3$, we need to target

$$\frac{800\ \text{cmil}}{A_{RMS}} \equiv \frac{800 \times 0.548 = 440\ \text{cmil}}{A_{COR}}$$

or

$$\frac{197{,}353}{800} \approx \frac{250\ \text{A}_{\text{RMS}}}{\text{cm}^2} \qquad \text{(in terms of RMS current)}$$

or

$$\frac{197{,}353}{440} = \frac{450\text{A}_{\text{COR}}}{\text{cm}^2} \qquad \text{(in terms of COR current, for } D = 0.3\text{)}$$

If the duty cycle was $D = 0.5$ (as in a Forward converter at lowest line condition), since $\sqrt{(0.5)} = 0.707$, we could write the target current density as

$$\frac{800\ \text{cmil}}{\text{A}_{\text{RMS}}} \equiv \frac{800 \times 0.707 = 565\ \text{cmil}}{\text{A}_{\text{COR}}}$$

or

$$\frac{197{,}353}{800} \approx \frac{250\ \text{A}_{\text{RMS}}}{\text{cm}^2} \qquad \text{(in terms of RMS current)}$$

or

$$\frac{197{,}353}{565} = \frac{350\ \text{A}_{\text{COR}}}{\text{cm}^2} \qquad \text{(in terms of COR current, for } D = 0.5\text{)}$$

We see that for both the above duty cycles, what remained constant was the following design target: a Forward converter transformer current density of 250 $A_{\text{RMS}}/\text{cm}^2$, exactly half the *widely, and blindly accepted* current density target. The underlying reason was F_R was at best 2, not 1.

We now recall our accurate equation for a Forward converter transformer:

$$\text{AP}_{\text{cu}_{P\_\text{cm}^4}} = \frac{645.49 \times P_{\text{IN}} \times J_{\text{cmil/A}_{\text{COR}}}}{f_{\text{Hz}} \times \Delta B_{\text{G}}} \quad \text{cm}^4$$

If we plug in our recommended current density of 800 cmil/A_{RMS}, i.e., 440 cmil/A_{COR} (for $D = 0.3$), and also assume that we have a utilization factor of 0.25 (ratio of primary winding area to core winding area, see Fig. 12.2), we get our basic recommendation to be

$$\text{AP}_{c_{\text{cm}^4}} = \frac{645.49 \times P_{\text{IN}} \times J_{\text{cmil/A}_{\text{COR}}}}{f_{\text{Hz}} \times \Delta B_{\text{G}}} = \frac{645.49 \times P_{\text{IN}} \times 440}{0.25 \times f_{\text{Hz}} \times \Delta B_{\text{G}}} = 11{,}360{,}624 \times \frac{P_{\text{IN}}}{f_{\text{Hz}} \times \Delta B_{\text{G}}}$$

or

$$\text{AP}_{c_{\text{cm}^4}} = 113.6 \times \frac{P_{\text{IN}}}{f_{\text{Hz}} \times \Delta B_{\text{T}}} \qquad \text{(Maniktala, for } D = 0.3,\ J = 250\ \text{A}_{\text{RMS}}/\text{cm}^2,\ K = 0.25\text{)}$$

Plugging in a typical value of $\Delta B = 1500$ G, we get

$$\text{AP}_{c_{\text{cm}^4}} = \frac{645.49 \times P_{\text{IN}} \times 440}{f_{\text{Hz}} \times 1500 \times 0.25} = 755 \times \frac{P_{\text{IN}}}{f_{\text{Hz}}}$$

Or equivalently (using kilohertz),

$$\text{AP}_{c_{\text{cm}^2}} = 0.75 \times \frac{P_{\text{IN}}}{f_{\text{kHz}}} \qquad \text{(Maniktala, for } D = 0.3,\ \Delta B = 0.15\ \text{T},\ J = 250\ \text{A}_{\text{RMS}}/\text{cm}^2,\ K = 0.25\text{)}$$

As we can see, this equation asks for a slightly larger core than we had suggested in the numerical example from the *A-Z* book. In the *A-Z* book, though, we had used a little more generous (conservative) current density, we also set a much more optimistic *utilization (fudge) factor*. We had derived

$$\text{AP}_{c_{\text{cm}^4}} = 675.6 \times \frac{P_{\text{IN}}}{f_{\text{Hz}}}$$

or equivalently

$$\text{AP}_{c_{\text{cm}^4}} = 0.676 \times \frac{P_{\text{IN}}}{f_{\text{kHz}}} \qquad \text{(Maniktala, for } D = 0.3,\ \Delta B = 0.15\ \text{T},\ J = 180\ \text{A}_{\text{RMS}}/\text{cm}^2,\ K = 0.38\text{)}$$

We conclude that the new equation we have now derived

$$AP_{c_{cm^2}} = 0.75 \times \frac{P_{IN}}{f_{kHz}} \qquad \text{(Maniktala, for } D = 0.3, \Delta B = 0.15\text{ T}, J = 250\text{ A}_{RMS}/\text{cm}^2, K = 0.25)$$

is a tad more realistic (and conservative in terms of available window area) than the older one in the *A-Z* book. This one asks for slightly higher area product (for a given power).

This slight modification of the *A-Z* book recommendation is a little more helpful for designing a safety-approved Class A Forward converter transformer running at a nominal $D = 0.3$.

Note that the underlying assumptions in our new equation include a *maximum flux swing of 1500 G, a current density of 250* A_{RMS}/cm^2*, and a utilization factor (ratio of primary winding area* W_{cu_p} *to the full core window area* W_{ac}*) of 0.25.*

If we have a core with a certain core-area product, we can also flip it to find its power capability as follows:

$$P_{IN} = \frac{AP_{c_{cm^4}} \times f_{Hz}}{754} = 1.33 \times 10^{-3} \times \left(AP_{c_{cm^4}} \times f_{Hz}\right)$$

$$P_{IN} = 1.33 \times AP_{c_{cm^4}} \times f_{kHz} \qquad \text{(Maniktala, for } D = 0.3, \Delta B = 0.15\text{ T},$$

$$J = 250\text{ A}_{RMS}/\text{cm}^2, K = 0.25)$$

For example, at f = 200 kHz, the ETD-34 core-set, with a core-area product of 1.66 cm^4, is suitable for

$$P_{IN} = \frac{1.66 \times 200000}{754} = 440\text{ W} \qquad \text{(recommendation example based on Maniktala)}$$

With an estimated efficiency of say 83 percent, this would work for a converter with $P_O = 365$ W.

Having understood this, we would like to compare with the equations others are espousing in related literature to see where we stand vis-à-vis their recommendations. Here is a list of other "similar" equations. It is a jungle where angels have feared to tread.

Industry-Recommended Equations for the Area Product of a Forward Converter

Fairchild Semi Recommendation

For example, see "The Forward-Converter Design Leverages Clever Magnetics by Carl Walding" at http://powerelectronics.com/mag/Fairchild.pdf:

$$AP_{mm^4} = \left(\frac{78.72 \times P_{IN}}{\Delta B \times f_{Hz}}\right)^{1.31} \times 10^4$$

This was alternatively expressed in Application Note AN-4134 from Fairchild as

$$AP_{mm^4} = \left(\frac{11.1 \times P_{IN}}{0.141 \times \Delta B \times f_{Hz}}\right)^{1.31} \times 10^4$$

But it is the same equation. It seems to be using the term area product, for the entire core. The field is in tesla. We can also rewrite this in terms of cm^4 as

$$AP_{c_{cm^4}} = \left(\frac{78.72 \times P_{IN}}{\Delta B_T \times f_{Hz}}\right)^{1.31} \qquad \text{(Fairchild)}$$

Compare it to our equation:

$$AP_{c_{cm^4}} = \frac{113.6 \times P_{IN}}{f_{Hz} \times \Delta B_T} \qquad \text{(Maniktala)}$$

We can simplify the Fairchild equation and set $\Delta B = 0.15$ T (the usual typical optimum flux swing to avoid saturation and keep core losses small). We get

$$AP_{c_{cm^4}} = \left(\frac{78.72 \times P_{IN}}{0.15 \times f_{kHz}}\right)^{1.31}$$

$$AP_{c_{cm^4}} = 0.43 \times \left(\frac{P_{IN}}{f_{kHz}}\right)^{1.31} \qquad \text{(Fairchild, with } \Delta B = 0.15\text{ T)}$$

We can compare this with our equation:

$$AP_{c_{cm^4}} = 0.75 \times \left(\frac{P_{IN}}{f_{kHz}}\right) \qquad \text{(Maniktala, with } \Delta B = 0.15\text{ T)}$$

For example, for 440-W input power, we know at 200 kHz, we recommend the ETD-34 with $AP_c = 1.66$ cm^4 (see Fig. 12.3). What does the Fairchild equation recommend? We get

$$AP_{c_{cm^4}} = 0.43 \times \left(\frac{440}{200}\right)^{1.31} = 1.21\text{ cm}^4 \qquad \text{(Fairchild recommendation example)}$$

ETD-29 has an area product (core) of 1.02 cm^4. So we will still end up using ETD-34. But, in general, at least for lower powers and frequencies, the Fairchild equation can ask for up to half the area product, thus implying much smaller cores. It seems more aggressive, and unless forced into a default larger core size, it will likely require either forced air cooling, or better (more expensive) core materials to compensate higher copper losses by much lower core losses. Or the transformer will be either non-safety-approved, or Class B safety-approved.

We can also solve the Fairchild equation for the power throughput from a given (core) area product (using typical $\Delta B = 1500$ G)

$$AP_{c_{cm^4}} = \left(\frac{78.72 \times P_{IN}}{0.15 \times f_{kHz}}\right)^{1.31}$$

$$P_{IN} = AP_{c_{cm^4}}^{0.763} \times \frac{0.15 \times f_{Hz}}{78.72} = 1.9 \times f_{kHz} \times AP_{c_{cm^4}}^{0.763}$$

$$P_{IN} = 1.9 \times f_{kHz} \times AP_{c_{cm^4}}^{0.763} \qquad \text{(Fairchild, for } \Delta B = 0.15\text{ T)}$$

TI/Unitrode Recommendation

For example, see www.ti.com/lit/ml/slup126/slup126.pdf and www.ti.com/lit/ml/slup205/slup205.pdf:

$$AP_{c_{cm^4}} = \left(\frac{11.1 \times P_{IN}}{K \times \Delta B_{Tesla} \times f_{Hz}}\right)^{1.143}$$

In this case K is a fudge factor related both to window utilization and topology. Unitrode asks to fix this at 0.141 for a single-ended Forward, and at 0.165 for a Bridge/half-Bridge. So with that, we get (for a single-ended Forward, assuming core area product as before):

$$AP_{c_{cm^4}} = \left(\frac{11.1 \times P_{IN}}{0.141 \times \Delta B_T \times f_{Hz}}\right)^{1.143} = \left(\frac{78.72 \times P_{IN}}{\Delta B_T \times f_{Hz}}\right)^{1.143} \qquad \text{(Unitrode)}$$

which is almost identical to the Fairchild equation, except that the exponent is mysteriously 1.143, leading to a much slower *rise* with power (and a *fall* with frequency), as compared to the exponent of 1.31 in the Fairchild equation. Note that this equation too (as the Fairchild equation) is said to be based on a current density of 450 A_{RMS}/cm^2—far more aggressive than the 250 A_{RMS}/cm^2 that we are espousing. But in all Unitrode application notes, the best achievable F_R was calculated to be just slightly larger than 1, *because it was based on sinusoidal waveforms*, whereas in reality, the best-case F_R is actually closer to 2, as we have assumed in our equations (see Fig. 11.8 too). That is why, our estimate seems excessively conservative, but is more accurate and realistic. However, in the TI/Unitrode recommendation, a better fit to actual data seems to have been created artificially, by introducing an arbitrary fudge factor K. Unfortunately, logically speaking, any such utilization factor should be changed or tweaked depending on the types of core being used. But that aspect is invariably overlooked.

We can also solve the Unitrode equation for the power throughput from a given (core) area product (using typical $\Delta B = 1500$ G)

$$P_{IN} = AP_{c_{cm^4}}^{0.875} \times \frac{0.15 \times f_{Hz}}{78.72} = 1.9 \times f_{kHz} \times AP_{c_{cm^4}}^{0.875}$$

$$P_{IN} = 1.9 \times f_{kHz} \times AP_{c_{cm^4}}^{0.875} \qquad \text{(Unitrode, for } \Delta B = 0.15\text{ T)}$$

Basso/On-Semi Recommendation

For example, see www.onsemi.com/pub_link/Collateral/TND350-D.PDF:

$$AP_{c_{cm^4}} = \left(\frac{P_O}{K \times \Delta B_T \times f_{Hz}}\right)^{4/3}$$

It is suggested that $K = 0.014$ for a Forward converter. This is another unexplained fudge factor really. Simplifying, we get for a Forward converter

$$AP_{c_{cm^4}} = \left(\frac{71.43 \times P_O}{\Delta B_T \times f_{Hz}}\right)^{1.33}$$

This is indeed very close to the Fairchild equation, though this equation unconsciously and implicitly assumes 100 percent efficiency, because it uses the output power instead of the input power, whereas, the worst-case assumption is that all the losses occur after passing through the transformer, not before (see Fig. 2.13). In that case, the transformer has to throughput the full input power, not the lower output power. To try and correct for this unfortunate assumption, we now assume 90 percent efficiency. We then get

$$AP_{c_{cm^4}} = \left(\frac{71.43 \times 0.9 \times P_{IN}}{\Delta B_T \times f_{Hz}}\right)^{1.33} = \left(\frac{64.3 \times P_{IN}}{\Delta B_{Tesla} \times f_{Hz}}\right)^{1.33} \qquad \text{(On-Semi corrected)}$$

Note that On-Semi says this is based on a window utilization factor of 0.4 and a current density of 420 A/cm^2. We assumed a 90 percent efficiency to get to the preceding equation.

The original, uncorrected On-Semi equation can also be written out for power throughput in terms of (core) area product as follows:

$$AP_{c_{cm^4}} = \left(\frac{71.43 \times P_O}{\Delta B_T \times f_{Hz}}\right)^{1.33} \Rightarrow \left(AP_{c_{cm^4}}\right)^{1/1.33} = \left(\frac{71.43 \times P_O}{\Delta B_T \times f_{Hz}}\right)$$

$$P_O = \frac{\Delta B_T \times f_{Hz}}{71.43} \times AP_{c_{cm^4}}^{0.752}$$

For a flux swing of 1500 G,

$$P_O = AP_{c_{cm^4}}^{0.752} \times \frac{0.15 \times f_{Hz}}{71.43} = 2.1 \times f_{kHz} \times AP_{c_{cm^4}}^{0.752}$$

$$P_{IN} = 2.1 \times AP_{c_{cm^4}}^{0.752} \times f_{kHz} \qquad \text{(On-Semi, for } \Delta B = 0.15\text{ T, 100 percent efficiency)}$$

ST Micro Recommendation

For example, see AN-1621 at http://www.st.com/st-web-ui/static/active/cn/resource/technical/document/application_note/CD00043746.pdf:

$$AP_{c_{cm^4}} = \left(\frac{67.2 \times P_{IN}}{\Delta B_T \times f_{Hz}}\right)^{1.31} \qquad \text{(ST Micro)}$$

$$P_{IN} = AP_{c_{cm^4}}^{0.763} \times \frac{0.15 \times f_{Hz}}{67.2} = 2.23 \times f_{kHz} \times AP_{c_{cm^4}}^{0.763}$$

$$P_{IN} = 2.23 \times f_{kHz} \times AP_{c_{cm^4}}^{0.763} \qquad \text{(ST Micro, for } \Delta B = 0.15\text{ T)}$$

Keith Billings and Pressman Recommendation and Explanation

For example, see *Switching Power Supply Design*, 3rd ed., by Abraham Pressman, Keith Billings, and Taylor Morey, and *Switchmode Power Supply Handbook* by Keith Billings.

Billing actually starts to derive the requisite equation in a manner identical to ours, based on basic principles, but then suddenly digresses and arrives at the exact same equation provided by TI/Unitrode previously, complete with the arbitrary fudge factor K.

This leads us to the origin of the odd exponent we are seeing in almost all the industrywide equations. Where did that come from? Almost all the equations are apparently based on an old empirical equation found in *Transformer and Inductor Design Handbook* by Colonel Wm. T. McLyman. The reason for the odd exponent stems from a *completely empirical statement* that says the optimum current density is not a constant as we assumed but is a function of area product. The paradox is that everyone (including Billings) continue to state the current density target as a fixed number anyway: 420 or 450 A/cm^2. But the inclusion of the odd exponent implies otherwise. Because, as indirectly explained by Billings himself in his derivation and his derivation parallels ours, except to the point that Billings plugs in McLyman's equation

$$J_{\mathrm{A/m^2}} = 450 \times 10^4 \times \mathrm{AP}^{-0.125}$$

So it seems that now the target current density is suddenly a function of area product, in direct contradiction to previous statements, which had declared the target to be a fixed value.

Nevertheless continuing the derivation as per Billings (ignoring fudge factors, etc., and replacing them with just an X here)

$$\mathrm{AP} = \frac{X \times P_{\mathrm{IN}}}{\mathrm{AP}^{-0.125} \times \Delta B \times f}$$

$$\mathrm{AP}^{1-0.125} = \mathrm{AP}^{0.875} = \frac{X \times P_{\mathrm{IN}}}{\Delta B \times f}$$

$$\mathrm{AP}^{0.875/0.875} = \mathrm{AP} = \left(\frac{X \times P_{\mathrm{IN}}}{\Delta B \times f}\right)^{1/0.875} = \left(\frac{X \times P_{\mathrm{IN}}}{\Delta B \times f}\right)^{1.143}$$

$$\mathrm{AP} = \left(\frac{X \times P_{\mathrm{IN}}}{\Delta B \times f}\right)^{1.143}$$

That is the underlying logic of how the strange exponent of 1.14 (or something else very close) appears in almost all equations, especially the early TI/Unitrode notes. Clearly, *the presence of this exponent implicitly assumes a variable current density, but that is not what is usually stated alongside*. Perhaps that explains the emergence of the fudge factors. It was just to get a better match to bench data. But, as mentioned, the fudge factors logically need to change with the transformer cores being used, and also their construction. For example, we may be using margin tape to comply with safety requirements, something that was ignored in the past. And so on. Besides, we realize that Dowell's equations, on which a lot of prior design equations seem to be based on, assumed sine waves. And so the AC resistance and transformer dissipation were severely under-estimated to start with. But two wrongs do not make a right.

Most prevalent equations seem to be far more aggressive at lower wattages than our recommendations. But it is possible that they will work. Keep in mind that since smaller transformers have a larger exposed surface area to volume, they cool better (smaller thermal resistance), and so inaccuracies in setting more aggressive current densities for smaller cores were perhaps not noticed, until larger cores appeared. In that case, temperatures rose much higher than expected. So now, empirically, it was decided to adjust the core size down for a given power requirement, just to get a larger surface area to allow it to cool, and of course a larger window area for allowing improved current density too. That is likely how the term –0.125 in the McLyman current density versus area product equation appeared, which in turn led to the odd exponents we see: such as 1.14, 1.31, and so on.

Disregarding where they all came from, we can certainly plot them all out for comparison to see if our guess about the historical sequence and the subsequent "equation adjustments," using fudge factors as described previously, seems plausible.

Plotting Industry Recommendations for Forward Converters

For a typical flux swing of 1500 G, we have plotted out the following recommendations:

$$P_{\mathrm{IN}} = 1.33 \times f_{\mathrm{kHz}} \times \mathrm{AP}_{c_{\mathrm{cm^4}}} \qquad \text{(Maniktala, see page 301)}$$

$$P_{\mathrm{IN}} = 1.9 \times f_{\mathrm{kHz}} \times \mathrm{AP}_{c_{\mathrm{cm^4}}}^{0.763} \qquad \text{(Fairchild, see page 302)}$$

$$P_{\mathrm{IN}} = 1.9 \times f_{\mathrm{kHz}} \times \mathrm{AP}_{c_{\mathrm{cm^4}}}^{0.875} \qquad \text{(Unitrode/TI, see page 302)}$$

$$P_{\mathrm{IN}} = 2.1 \times f_{\mathrm{kHz}} \times \mathrm{AP}_{c_{\mathrm{cm^4}}}^{0.752} \qquad \text{(On-Semi, see page 303)}$$

$$P_{\mathrm{IN}} = 2.23 \times f_{\mathrm{kHz}} \times \mathrm{AP}_{c_{\mathrm{cm^4}}}^{0.763} \qquad \text{(ST Micro, see page 303)}$$

We see from these that, indeed, doubling the frequency will double the power (so we really do not need to plot out curves for 300 kHz, 400 kHz, and so on—it is obvious how to derive the results for different frequencies).

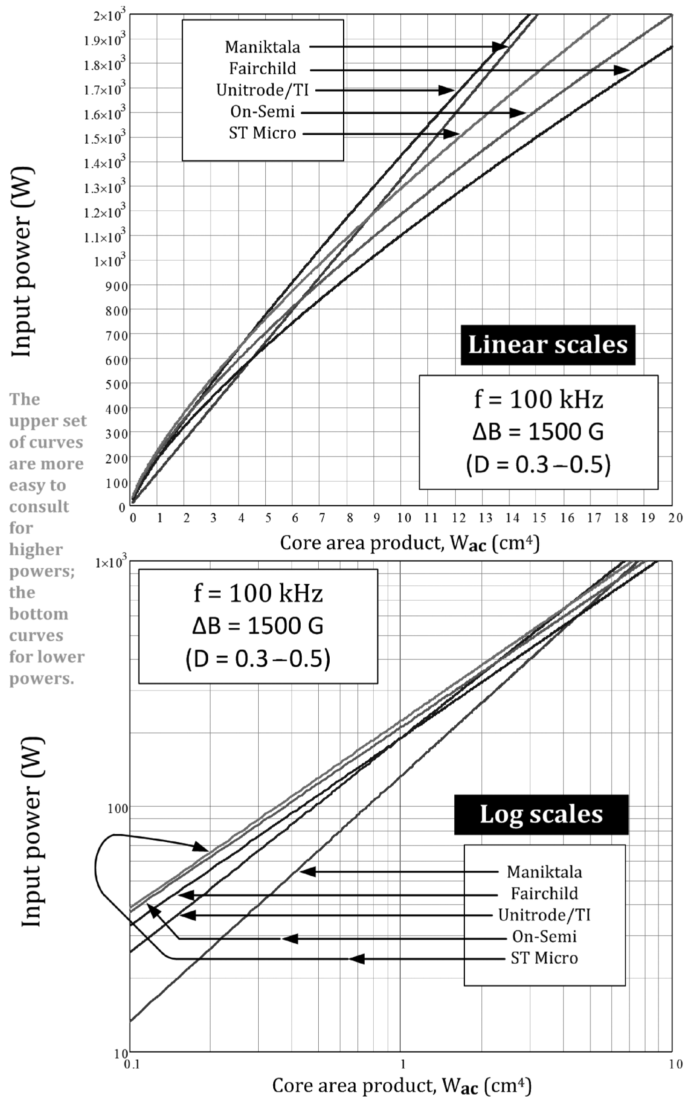


Figure 12.4 Comparing industry recommendations through plots of power versus core area product, assuming typical flux swing of 1500 G (at 100 kHz).

On plotting these out in Figs. 12.4 and 12.5, we see that our recommendation is more conservative for smaller output powers but is in line with others at higher power levels. We know that ours is consistently based on a constant current density target of 250 A_{RMS}/cm^2. The other recommendations do seem to be using a variable-current density target, though that is never explicitly defined in literature. They may *get away* with their more aggressive core-size recommendations *for small cores*, based on the empirical fact that smaller cores have improved thermal resistances on the bench, because of their higher ratio of surface-area to volume. And that fact may admittedly allow us also to judiciously increase the current density in small cores, say up to 350 to 400 A_{RMS}/cm^2. But it is quite clear that for larger cores, we do need to drop down to 250 A_{RMS}/cm^2 because all other recommendations do coincide with ours at high power levels, and our recommendation was based on a fixed 250 A_{RMS}/cm^2.

We can confirm from Fig. 12.5 that our recommendation is ETD34 ($AP_c = 1.66$ cm^4) for up to 440-W input power at 200 kHz, whereas the others typically allow 100 to 200 W more than that.

We can also compare with another set of curves historically available from Magnetics® at www.mag-inc.com. These are shown in Fig. 12.6 and are clearly the most aggressive. They also do not seem to spell out clearly if the topology is a single-ended Forward converter, or say, a Push-Pull (where due to symmetric excitation, some engineers claim it will give exactly twice the power reflected by the curves in Figs. 12.4 and 12.5). Keep in mind that the Magnetics Inc. curves seem to be based on low-frequency sine waves applied to test cores. Yet they were widely "referred to" in most of the prevailing Forward converter design notes around us even today.

Our conclusion is the equations proposed by us are self-consistent, derived from first principles, and less likely to run into thermally-initiated recalls.

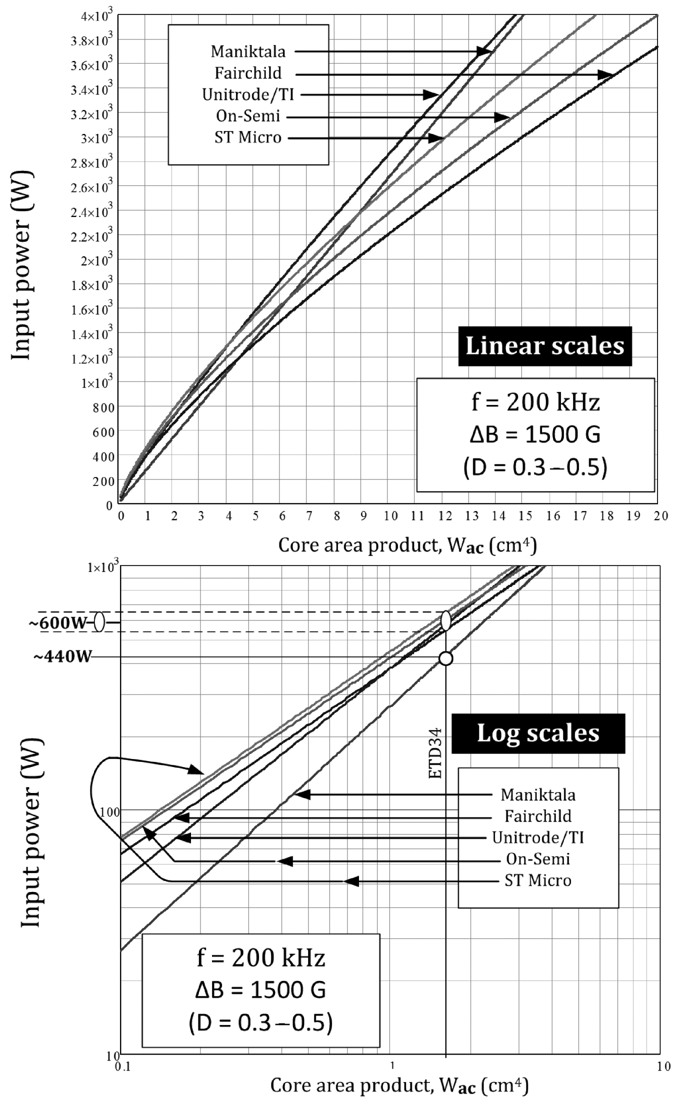


Figure 12.5 Comparing industry recommendations through plots of power versus core area product, assuming typical flux swing of 1500 G (at 200 kHz).

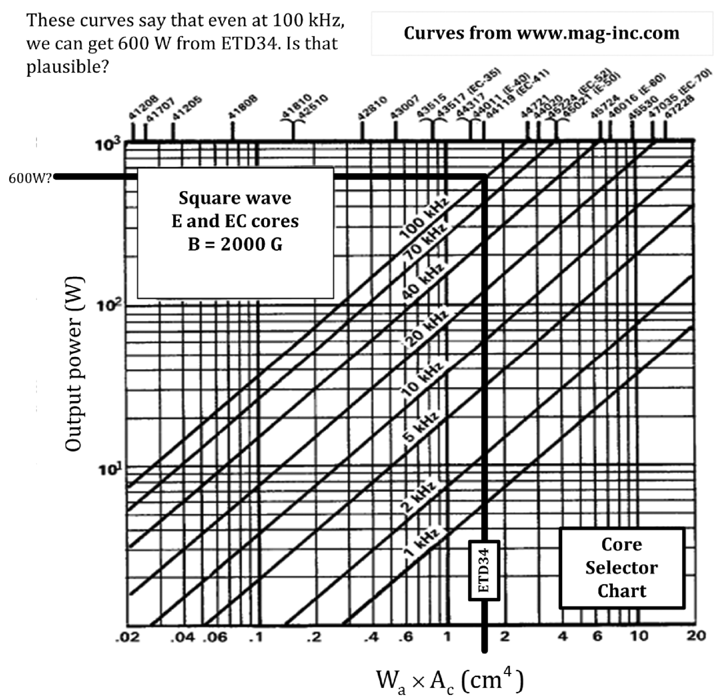


Figure 12.6 Historically available recommendations from Magnetics Inc.

Area Product for Symmetric Converters

We derived the general equation for a Forward converter

$$\Delta B_G = \frac{P_{IN}}{(J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

The following similar equation is available at www.cedt.iisc.ernet.in/people//lums/smpc_page/pes03/pes03.pdf from Dr. Umanand at CEDT, Bangalore. This is derived in a manner similar to ours, but it does not use the COR current density to *mask* the dependence on D. Instead it uses the RMS current density and brings out $\sqrt{D}$ explicitly.

$$A_P = \frac{\sqrt{D_{max}}\, P_O (1 + 1/\eta)}{K_w J B_m f_s}$$

The paper also provides expressions for half-bridge, full-bridge, and Push-Pull converters. Note that it includes the efficiency η too. But we are setting that equal to 1 here (and simply changing P_O to P_{IN}). This is to retain simplicity. Also, casting the equations in terms of our ongoing nomenclature and units, we can rewrite the applicable equations as shown subsequently. Note that we have continued to include the correction factor of 0.785 for round wires so that the current density explicitly refers to the current passing through copper—though for foil windings we can actually remove it, but we will ignore that fact in our discussions for simplicity too. Note that now the current density is expressed explicitly in terms of RMS current, not the COR current value. We have

$$\Delta B_G = \frac{P_{IN}\sqrt{D_{MAX}}}{(J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8 \qquad \text{(Forward, RMS current density)}$$

For $D = 0.5$, we get

$$\Delta B_G = \frac{P_{IN} \times 0.707}{(J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8 \qquad \text{(Forward, RMS current density)}$$

Simplifying

$$\Delta B_G = \frac{0.90 \times P_{IN}}{(J_{A/cm^2}) \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8 \qquad \text{(Forward, RMS current density)}$$

For half-bridge and full-bridge converters we have

$$\Delta B_G = \frac{P_{IN}(1 + \sqrt{2})}{4 \times (J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

$$= \frac{0.6 \times P_{IN}}{(J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

$$= \frac{0.764 \times P_{IN}}{(J_{A/cm^2}) \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8 \quad \text{(Half- or full-bridge, RMS current density)}$$

We see that this says, in effect, for the same area product, flux swing, etc., the power throughput of the half-bridge and Push-Pull converter is greater than that for a single-ended Forward converter by the factor $0.9/0.764 = 1.18$, that is, only 18 percent more, unless we increase the flux swing, assuming that is acceptable in terms of core loss. More on that soon.

For a Push-Pull converter we have

$$\Delta B_G = \frac{P_{IN}}{\sqrt{2} \times (J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

$$= \frac{0.71 \times P_{IN}}{(J_{A/cm^2}) \times 0.785 \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8$$

$$= \frac{0.90 \times P_{IN}}{(J_{A/cm^2}) \times AP_{cu_{P\_cm^4}} \times f_{Hz}} \times 10^8 \qquad \text{(Push-Pull, RMS current density)}$$

which is the same as the Forward converter at $D = 0.5$! This requires some explanation.

Historically, the flux swing was restricted to 1500 G in a single-ended Forward converter, because it was known that the ferrite core could saturate at about 3000 G. So to avoid saturation during sudden line and load transients, a headroom of 1500 G was maintained. When engineers set to work on the Push-Pull, half-bridge, and full-bridge converters, since the core excitation was symmetric (around 0 G), the total flux swing could be increased to 3000 G (± 1500 G), and we would still have the same 1500-G safety margin. So at first sight it was felt that we could double the power throughput in most cases. It is perhaps still possible, but only at low switching frequencies (around 20 kHz).

Today, at higher switching frequencies, the flux swing is kept to 1500 G to keep core loss down to 100 mW/cm^3 (for 3F3 material at 200 kHz, for example). The response of current limit circuitry, etc., is fast enough to minimize worries about hitting B_{SAT} under transients. So, in fact, *a good design (from a thermal viewpoint) may even set ΔB to 1000 or 1200 G only.* In other words, even in a *symmetric excitation converter*, we would very likely continue to keep ΔB to less than 1500 G. In that case, the preceding discussions and equations (derived from first principles), tell us there is very little to gain in terms of reduction of size of magnetic components in moving from a single-ended Forward converter to a symmetric converter. Yes, doing so may greatly help in finding components to ensure high efficiencies at high powers, and so on. But the usual *rule of thumb* that we can blindly double the power from a given core by using, say, the half-bridge instead of a Forward, is very doubtful indeed—from the thermal viewpoint in particular.

More Accurate Estimate of Power Throughput in Safety Transformers

All recommendations so far have been based on an assumption of a certain window utilization factor. All the curves we have shown in Figs. 12.4 and 12.5 have some such underlying assumption. At least in our case we have rather clearly assumed (and announced) that the primary windings will occupy exactly one-fourth of the total available *core* window area (i.e., $K = 0.25$). Most others typically provide rather vague utilization numbers, seemingly applied to somehow fit empirical data, but usually provide almost no physical explanation.

We also opined that for smaller transformers, we may be able to target higher current densities judiciously. Keep in mind that if the (exposed) area of a core was proportional to its volume, then even assuming that the coefficient of convection h was constant with respect to area (it isn't perfectly), we would expect the thermal resistance, which is assumed inversely proportional to surface area, to be inversely proportional to the volume (size of core) too. So, we would expect Rth to vary as per $1/V_e$. But that does not happen. The actual thermal resistance is much *worse than expected* for larger cores and is based on the following well-known empirical formula. See Fig. 12.7 for how a "wishful situation" was tempered with reality. So, the accepted empirical equation is

$$\text{Rth} = \frac{53}{V_e^{0.54}} \quad °\text{C/W}$$

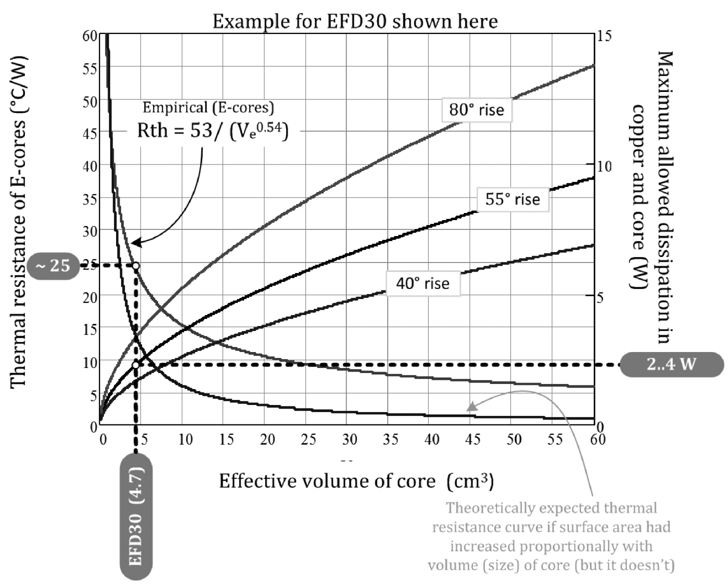


Figure 12.7 Thermal resistance of E-cores and maximum allowed dissipation (in windings and core).

However, we should also keep in mind that in smaller cores, less and less window utilization occurs, because the margin tape is of a fixed width (and also with a constant bobbin wall thickness), and does not decrease proportionally with core window area. So we will likely struggle even to maintain the same fixed current density. We just may not have enough winding width available, once we subtract the margin tape width on either side.

To more accurately judge what is the *real utilization factor* to plug in (instead of the default value of 0.25 we have used so far), *we need to actually compute the physical dimensions*, making some assumptions about bobbin wall thickness too. We start with some popular core sizes listed in Table 12.1, and then use that to arrive at the detailed results in Tables 12.2 to 12.5, cranked out by a spreadsheet, for the following cases: no margin tape, 2-mm margin tape (telecommunication applications), 4-mm margin tape (AC-DC with no PFC), 6.3-mm margin tape (AC-DC with Boost PFC front-end). As we can see, certain core sizes result in "NA" (nonapplicable), because after subtracting the margin tape from the available bobbin width, we get either almost no space for any winding, or worse, we have negative space. We also see that the utilization factor K_{cu_p}, is all over the place. Even our assumption of $K = 0.25$ was clearly a broad assumption, not really valid for small cores in particular. From these tables we can do a much more detailed and accurate calculation, as we will carry out shortly.

Basic Core Parameters (see Fig. 12.2)									
***A* (mm)**	***B* (mm)**	***C* (mm)**	***D* (mm)**	***E* (mm)**	***F* (mm)**	**l_e (cm)**	**A_e (cm^2)**	**V_e (cm^3)**	**Core**
20.00	10	5	6.3	12.8	5.2	4.28	0.312	1.34	**EE20/10/5**
25.00	10	6	6.4	18.8	6.35	4.9	0.395	1.93	**EE25/10/6**
35.00	18	10	12.5	24.5	10	8.07	1	8.07	**EE35/18/10**
42.00	21	15	14.8	29.5	12.2	9.7	1.78	17.3	**EE42/21/15**
42.00	21	20	14.8	29.5	12.2	9.7	2.33	22.7	**EE42/21/20**
55.00	28	20	18.5	37.5	17.2	12.3	4.2	52	**EE55/28/20**
28.00	14	11	9.75	21.75	9.9	6.4	0.814	5.26	**ER28/14/11**
35.00	20.7	11.3	14.7	25.6	11.3	9.08	1.07	9.72	**ER35/21/11**
42.00	22	16	15.45	30.05	15.5	9.88	1.94	19.2	**ER42/22/16**
54.00	18	18	11.1	40.65	17.9	9.18	2.5	23	**ER54/18/18**
12.00	6	3.5	4.55	9	5.4	2.85	0.114	0.325	**EFD12/6/3.5**
15.00	8	5	5.5	11	5.3	3.4	0.15	0.51	**EFD15/8/5**
20.00	10	7	7.7	15.4	8.9	4.7	0.31	1.46	**EFD20/10/7**
25.00	13	9	9.3	18.7	11.4	5.7	0.58	3.3	**EFD25/13/9**
30.00	15	9	11.2	22.4	14.6	6.8	0.69	4.7	**EFD30/15/9**
29.00	16	10	11	22	9.8	7.2	0.76	5.47	**ETD29/16/10**
34.00	17	11	11.8	25.6	11.1	7.86	0.97	7.64	**ETD34/17/11**
39.00	20	13	14.2	29.3	12.8	9.22	1.25	11.5	**ETD39/20/13**
44.00	22	15	16.1	32.5	15.2	10.3	1.73	17.8	**ETD44/22/15**
49.00	25	16	17.7	36.1	16.7	11.4	2.11	24	**ETD49/25/16**
54.00	28	19	20.2	41.2	18.9	12.7	2.8	35.5	**ETD54/28/19**
59.00	31	22	22.5	44.7	21.65	13.9	3.68	51.5	**ETD59/31/22**
74.00	29.5	NA	20.35	57.5	29.5	12.8	7.9	101	**PM74/59**
87.00	35	NA	24	67	31.7	14.6	9.1	133	**PM87/70**
114.00	46.5	NA	31.5	88	43	20	17.2	344	**PM114/93**
35.00	17.3	9.5	12.3	22.75	9.5	7.74	0.843	6.53	**EC35**
41.00	19.5	11.6	13.9	27.05	11.6	8.93	1.21	10.8	**EC41**
52.00	24.2	13.4	15.9	33	13.4	10.5	1.8	18.8	**EC52**
70.00	34.5	16.4	22.75	44.5	16.4	14.4	2.79	40.1	**EC70**

Table 12.1 Selection of Popular Cores with Basic Characteristics

0-mm (no) margin tape on either side.
Default values: 1.15-mm bobbin wall along direction of *A*, 1.35-mm bobbin wall along direction of *D*, additional 0.35-mm minimum clearance to the ferrite on the outside of the copper winding. See Fig. 12.2.

W_{ac} (cm²)	W_{ab} (cm²)	Width (mm)	Height (mm)	AP_b (cm⁴)	AP_c (cm⁴)	Width _tape (mm)	W_{cu} (cm²)	AP_{cu_P} (cm⁴)	K_{cu_P}	MLT (cm)	Core
0.48	0.23	9.90	2.30	0.07	0.15	9.90	0.23	0.04	0.24	4.02	**EE20/10/5**
0.80	0.48	10.10	4.73	0.19	0.31	10.10	0.48	0.09	0.30	5.42	**EE25/10/6**
1.81	1.28	22.30	5.75	1.28	1.81	22.30	1.28	0.64	0.35	7.36	**EE35/18/10**
2.56	1.92	26.90	7.15	3.42	4.56	26.90	1.92	1.71	0.38	9.36	**EE42/21/15**
2.56	1.92	26.90	7.15	4.48	5.97	26.90	1.92	2.24	0.38	10.36	**EE42/21/20**
3.76	2.97	34.30	8.65	12.46	15.77	34.30	2.97	6.23	0.40	11.96	**EE55/28/20**
1.16	0.74	16.80	4.43	0.61	0.94	16.80	0.74	0.30	0.32	5.33	**ER28/14/11**
2.10	1.51	26.70	5.65	1.61	2.25	26.70	1.51	0.81	0.36	6.16	**ER35/21/11**
2.25	1.63	28.20	5.78	3.16	4.36	28.20	1.63	1.58	0.36	7.52	**ER42/22/16**
2.53	1.93	19.50	9.88	4.81	6.31	19.50	1.93	2.41	0.38	9.56	**ER54/18/18**
0.16	0.02	6.40	0.30	0.00	0.02	6.40	0.02	0.00	0.06	2.68	**EFD12/6/3.5**
0.31	0.11	8.30	1.35	0.02	0.05	8.30	0.11	0.01	0.18	3.23	**EFD15/8/5**
0.50	0.22	12.70	1.75	0.07	0.16	12.70	0.22	0.03	0.22	4.24	**EFD20/10/7**
0.68	0.34	15.90	2.15	0.20	0.39	15.90	0.34	0.10	0.25	5.22	**EFD25/13/9**
0.87	0.47	19.70	2.40	0.33	0.60	19.70	0.47	0.16	0.27	5.89	**EFD30/15/9**
1.34	0.89	19.30	4.60	0.67	1.02	19.30	0.89	0.34	0.33	5.36	**ETD29/16/10**
1.71	1.20	20.90	5.75	1.17	1.66	20.90	1.20	0.58	0.35	6.13	**ETD34/17/11**
2.34	1.73	25.70	6.75	2.17	2.93	25.70	1.73	1.08	0.37	6.97	**ETD39/20/13**
2.79	2.11	29.50	7.15	3.65	4.82	29.50	2.11	1.82	0.38	7.85	**ETD44/22/15**
3.43	2.68	32.70	8.20	5.66	7.25	32.70	2.68	2.83	0.39	8.66	**ETD49/25/16**
4.50	3.64	37.70	9.65	10.19	12.61	37.70	3.64	5.09	0.40	9.80	**ETD54/28/19**
5.19	4.24	42.30	10.03	15.61	19.09	42.30	4.24	7.80	0.41	10.78	**ETD59/31/22**
5.70	4.75	38.00	12.50	37.53	45.01	38.00	4.75	18.76	0.42	14.03	**PM74/59**
8.47	7.32	45.30	16.15	66.58	77.10	45.30	7.32	33.29	0.43	15.87	**PM87/70**
14.18	12.66	60.30	21.00	217.80	243.81	60.30	12.66	108.90	0.45	20.94	**PM114/93**
1.63	1.12	21.90	5.13	0.95	1.37	21.90	1.12	0.47	0.34	5.43	**EC35**
2.15	1.56	25.10	6.23	1.89	2.60	25.10	1.56	0.95	0.36	6.43	**EC41**
3.12	2.42	29.10	8.30	4.35	5.61	29.10	2.42	2.17	0.39	7.65	**EC52**
6.39	5.37	42.80	12.55	14.99	17.84	42.80	5.37	7.49	0.42	9.93	**EC70**

W_{ac} is window area of core; W_{ab} is window area in side bobbin; **Width** is the width of any layer inside bobbin if no margin tape were present; **Height** is the height available for winding copper; $\mathbf{AP}_b$ is the area product of the bobbin; $\mathbf{AP}_c$ is the area product of the core; **Width_tape** is the actual width available for the copper layer with margin tape present; W_{cu} is the net window area available to wind copper (in Primary and Secondary) with margin tape and bobbin considered; $\mathbf{AP}_{cu_P}$ is the area product available for primary winding alone, assuming it is half the total available; K_{cu_P} is the actual utilization factor for the primary winding (ratio of AP_{cu_P} to AP_c), **MLT** is the mean (or average) length per turn with the bobbin wall thickness and required minimum clearance considered.

TABLE 12.2 Popular Cores with Area Product, Window Area, Utilization Factor with No Margin Tape

Number of Primary Turns

This is another source of confusion. Most databooks, from core vendors in particular, ask to fix the number of primary turns based on the equation for square waves shown in Fig. 12.8. Many engineers use that as a basis but do not realize that it uses the RMS value of the voltage, not the input DC rail. Besides, it assumes 50 percent duty cycle as we can see from the derivation in the figure too. We therefore do not recommend using it. The correct relationship must involve the duty cycle, just as we concluded during the core selection process too.

2-mm margin tape on either side.
Default values: 1.15-mm bobbin wall along direction of *A*, 1.35-mm bobbin wall along direction of *D*, additional 0.35-mm minimum clearance to the ferrite on the outside of the copper winding. See Fig. 12.2.

W_{ac} (cm²)	W_{ab} (cm²)	Width (mm)	Height (mm)	AP_b (cm⁴)	AP_c (cm⁴)	Width _tape (mm)	W_{cu} (cm²)	AP_{cu_P} (cm⁴)	K_{cu_P}	MLT (cm)	Core
0.48	0.23	9.90	2.30	0.07	0.15	5.90	0.14	0.02	0.14	4.02	**EE20/10/5**
0.80	0.48	10.10	4.73	0.19	0.31	6.10	0.29	0.06	0.18	5.42	**EE25/10/6**
1.81	1.28	22.30	5.75	1.28	1.81	18.30	1.05	0.53	0.29	7.36	**EE35/18/10**
2.56	1.92	26.90	7.15	3.42	4.56	22.90	1.64	1.46	0.32	9.36	**EE42/21/15**
2.56	1.92	26.90	7.15	4.48	5.97	22.90	1.64	1.91	0.32	10.36	**EE42/21/20**
3.76	2.97	34.30	8.65	12.46	15.77	30.30	2.62	5.50	0.35	11.96	**EE55/28/20**
1.16	0.74	16.80	4.43	0.61	0.94	12.80	0.57	0.23	0.25	5.33	**ER28/14/11**
2.10	1.51	26.70	5.65	1.61	2.25	22.70	1.28	0.69	0.31	6.16	**ER35/21/11**
2.25	1.63	28.20	5.78	3.16	4.36	24.20	1.40	1.36	0.31	7.52	**ER42/22/16**
2.53	1.93	19.50	9.88	4.81	6.31	15.50	1.53	1.91	0.30	9.56	**ER54/18/18**
0.16	0.02	6.40	0.30	0.00	0.02	2.40	0.01	0.00	0.02	2.68	**EFD12/6/3.5**
0.31	0.11	8.30	1.35	0.02	0.05	4.30	0.06	0.00	0.09	3.23	**EFD15/8/5**
0.50	0.22	12.70	1.75	0.07	0.16	8.70	0.15	0.02	0.15	4.24	**EFD20/10/7**
0.68	0.34	15.90	2.15	0.20	0.39	11.90	0.26	0.07	0.19	5.22	**EFD25/13/9**
0.87	0.47	19.70	2.40	0.33	0.60	15.70	0.38	0.13	0.22	5.89	**EFD30/15/9**
1.34	0.89	19.30	4.60	0.67	1.02	15.30	0.70	0.27	0.26	5.36	**ETD29/16/10**
1.71	1.20	20.90	5.75	1.17	1.66	16.90	0.97	0.47	0.28	6.13	**ETD34/17/11**
2.34	1.73	25.70	6.75	2.17	2.93	21.70	1.46	0.92	0.31	6.97	**ETD39/20/13**
2.79	2.11	29.50	7.15	3.65	4.82	25.50	1.82	1.58	0.33	7.85	**ETD44/22/15**
3.43	2.68	32.70	8.20	5.66	7.25	28.70	2.35	2.48	0.34	8.66	**ETD49/25/16**
4.50	3.64	37.70	9.65	10.19	12.61	33.70	3.25	4.55	0.36	9.80	**ETD54/28/19**
5.19	4.24	42.30	10.03	15.61	19.09	38.30	3.84	7.06	0.37	10.78	**ETD59/31/22**
5.70	4.75	38.00	12.50	37.53	45.01	34.00	4.25	16.79	0.37	14.03	**PM74/59**
8.47	7.32	45.30	16.15	66.58	77.10	41.30	6.67	30.35	0.39	15.87	**PM87/70**
14.18	12.66	60.30	21.00	217.80	243.81	56.30	11.82	101.68	0.42	20.94	**PM114/93**
1.63	1.12	21.90	5.13	0.95	1.37	17.90	0.92	0.39	0.28	5.43	**EC35**
2.15	1.56	25.10	6.23	1.89	2.60	21.10	1.31	0.79	0.31	6.43	**EC41**
3.12	2.42	29.10	8.30	4.35	5.61	25.10	2.08	1.87	0.33	7.65	**EC52**
6.39	5.37	42.80	12.55	14.99	17.84	38.80	4.87	6.79	0.38	9.93	**EC70**

W_{ac} is window area of core; W_{ab} is window area in side bobbin; **Width** is the width of any layer inside bobbin if no margin tape were present; **Height** is the height available for winding copper; AP_b is the area product of the bobbin; AP_c is the area product of the core; **Width_tape** is the actual width available for the copper layer with margin tape present; W_{cu} is the net window area available to wind copper (in Primary and Secondary) with margin tape and bobbin considered; AP_{cu_P} is the area product available for primary winding alone, assuming it is half the total available; K_{cu_P} is the actual utilization factor for the primary winding (ratio of AP_{cu_P} to AP_c), **MLT** is the mean (or average) length per turn with the bobbin wall thickness and required minimum clearance considered.

Table 12.3 Popular Cores with Area Product, Window Area, Utilization Factor with 2-mm Margin Tape

Other engineers (such as AN-4134 from Fairchild) ask to use this equation:

$$N_{P\_MIN} = \frac{V_{INMIN} \times D_{MAX}}{A_e \times f \times \Delta B} \times 10^6$$

We need to correct some wrong impressions here. There is actually no need to do the calculation at minimum input. The reason is that the duty cycle of a Forward converter is based on the Buck cell that follows the transformer stage, which has an effective DC input of V_{INR} (the reflected input voltage) and an output of V_O. So we have (as for a Buck)

$$D = \frac{V_O}{V_{INR}} = \frac{n \times V_O}{V_{IN}}$$

4-mm margin tape on either side.
Default values: 1.15-mm bobbin wall along direction of *A*, 1.35-mm bobbin wall along direction of *D*, additional 0.35-mm minimum clearance to the ferrite on the outside of the copper winding. See Fig. 12.2.

W_{ac} (cm²)	W_{ab} (cm²)	Width (mm)	Height (mm)	AP_b (cm⁴)	AP_c (cm⁴)	Width _tape (mm)	W_{cu} (cm²)	AP_{cu_P} (cm⁴)	K_{cu_P}	MLT (cm)	Core
0.48	0.23	9.90	2.30	0.07	0.15	1.90	NA	NA	0.05	4.02	**EE20/10/5**
0.80	0.48	10.10	4.73	0.19	0.31	2.10	NA	NA	NA	5.42	**EE25/10/6**
1.81	1.28	22.30	5.75	1.28	1.81	14.30	0.82	0.41	0.23	7.36	**EE35/18/10**
2.56	1.92	26.90	7.15	3.42	4.56	18.90	1.35	1.20	0.26	9.36	**EE42/21/15**
2.56	1.92	26.90	7.15	4.48	5.97	18.90	1.35	1.57	0.26	10.36	**EE42/21/20**
3.76	2.97	34.30	8.65	12.46	15.77	26.30	2.27	4.78	0.30	11.96	**EE55/28/20**
1.16	0.74	16.80	4.43	0.61	0.94	8.80	0.39	0.16	0.17	5.33	**ER28/14/11**
2.10	1.51	26.70	5.65	1.61	2.25	18.70	1.06	0.57	0.25	6.16	**ER35/21/11**
2.25	1.63	28.20	5.78	3.16	4.36	20.20	1.17	1.13	0.26	7.52	**ER42/22/16**
2.53	1.93	19.50	9.88	4.81	6.31	11.50	1.14	1.42	0.22	9.56	**ER54/18/18**
0.16	0.02	6.40	0.30	0.00	0.02	NA	NA	NA	NA	2.68	**EFD12/6/3.5**
0.31	0.11	8.30	1.35	0.02	0.05	0.30	NA	NA	NA	3.23	**EFD15/8/5**
0.50	0.22	12.70	1.75	0.07	NA	NA	NA	NA	NA	4.24	**EFD20/10/7**
0.68	0.34	15.90	2.15	0.20	0.39	7.90	0.17	0.05	0.13	5.22	**EFD25/13/9**
0.87	0.47	19.70	2.40	0.33	0.60	11.70	0.28	0.10	0.16	5.89	**EFD30/15/9**
1.34	0.89	19.30	4.60	0.67	1.02	11.30	0.52	0.20	0.19	5.36	**ETD29/16/10**
1.71	1.20	20.90	5.75	1.17	1.66	12.90	0.74	0.36	0.22	6.13	**ETD34/17/11**
2.34	1.73	25.70	6.75	2.17	2.93	17.70	1.19	0.75	0.25	6.97	**ETD39/20/13**
2.79	2.11	29.50	7.15	3.65	4.82	21.50	1.54	1.33	0.28	7.85	**ETD44/22/15**
3.43	2.68	32.70	8.20	5.66	7.25	24.70	2.03	2.14	0.29	8.66	**ETD49/25/16**
4.50	3.64	37.70	9.65	10.19	12.61	29.70	2.87	4.01	0.32	9.80	**ETD54/28/19**
5.19	4.24	42.30	10.03	15.61	19.09	34.30	3.44	6.33	0.33	10.78	**ETD59/31/22**
5.70	4.75	38.00	12.50	37.53	45.01	30.00	3.75	14.81	0.33	14.03	**PM74/59**
8.47	7.32	45.30	16.15	66.58	77.10	37.30	6.02	27.41	0.36	15.87	**PM87/70**
14.18	12.66	60.30	21.00	217.80	243.81	52.30	10.98	94.45	0.39	20.94	**PM114/93**
1.63	1.12	21.90	5.13	0.95	1.37	13.90	0.71	0.30	0.22	5.43	**EC35**
2.15	1.56	25.10	6.23	1.89	2.60	17.10	1.06	0.64	0.25	6.43	**EC41**
3.12	2.42	29.10	8.30	4.35	5.61	21.10	1.75	1.58	0.28	7.65	**EC52**
6.39	5.37	42.80	12.55	14.99	17.84	34.80	4.37	6.09	0.34	9.93	**EC70**

W_{ac} is window area of core; W_{ab} is window area in side bobbin; **Width** is the width of any layer inside bobbin if no margin tape were present; **Height** is the height available for winding copper; $\mathbf{AP}_b$ is the area product of the bobbin; $\mathbf{AP}_c$ is the area product of the core; **Width_tape** is the actual width available for the copper layer with margin tape present; W_{cu} is the net window area available to wind copper (in Primary and Secondary) with margin tape and bobbin considered; $\mathbf{AP}_{cu_P}$ is the area product available for primary winding alone, assuming it is half the total available; $\mathbf{K}_{cu_P}$ is the actual utilization factor for the primary winding (ratio of AP_{cu_P} to AP_c), **MLT** is the mean (or average) length per turn with the bobbin wall thickness and required minimum clearance considered.

Table 12.4 Popular Cores with Area Product, Window Area, Utilization Factor with 4-mm Margin Tape

where $n = N_P/N_S$. The volt-seconds across the primary winding is

$$\text{Volt-seconds} = V_{\text{IN}} \times \frac{D}{f}$$

So at minimum input we get

$$\text{Volt-seconds}_{\text{MIN}} = V_{\text{INMIN}} \times \frac{1}{f} \times \frac{N_P \times V_O}{V_{\text{INMIN}}}$$

6.3-mm margin tape on either side.
Default values: 1.15-mm bobbin wall along direction of *A*, 1.35-mm bobbin wall along direction of *D*, additional 0.35-mm minimum clearance to the ferrite on the outside of the copper winding. See Fig. 12.2.

W_{ac} (cm²)	W_{ab} (cm²)	Width (mm)	Height (mm)	AP_b (cm⁴)	AP_c (cm⁴)	Width _tape (mm)	W_{cu} (cm²)	AP_{cu_P} (cm⁴)	K_{cu_P}	MLT (cm)	Core
0.48	0.23	9.90	2.30	0.07	0.15	NA	NA	NA	NA	4.02	**EE20/10/5**
0.80	0.48	10.10	4.73	0.19	0.31	NA	NA	NA	NA	5.42	**EE25/10/6**
1.81	1.28	22.30	5.75	1.28	1.81	9.70	0.56	0.28	0.15	7.36	**EE35/18/10**
2.56	1.92	26.90	7.15	3.42	4.56	14.30	1.02	0.91	0.20	9.36	**EE42/21/15**
2.56	1.92	26.90	7.15	4.48	5.97	14.30	1.02	1.19	0.20	10.36	**EE42/21/20**
3.76	2.97	34.30	8.65	12.46	15.77	21.70	1.88	3.94	0.25	11.96	**EE55/28/20**
1.16	0.74	16.80	4.43	0.61	0.94	4.20	0.19	0.08	0.08	5.33	**ER28/14/11**
2.10	1.51	26.70	5.65	1.61	2.25	14.10	0.80	0.43	0.19	6.16	**ER35/21/11**
2.25	1.63	28.20	5.78	3.16	4.36	15.60	0.90	0.87	0.20	7.52	**ER42/22/16**
2.53	1.93	19.50	9.88	4.81	6.31	6.90	0.68	0.85	0.13	9.56	**ER54/18/18**
0.16	0.02	6.40	0.30	0.00	0.02	NA	NA	NA	NA	2.68	**EFD12/6/3.5**
0.31	0.11	8.30	1.35	0.02	0.05	NA	NA	NA	NA	3.23	**EFD15/8/5**
0.50	0.22	12.70	1.75	0.07	0.16	0.10	NA	NA	NA	4.24	**EFD20/10/7**
0.68	0.34	15.90	2.15	0.20	0.39	3.30	0.07	0.02	0.05	5.22	**EFD25/13/9**
0.87	0.47	19.70	2.40	0.33	0.60	7.10	0.17	0.06	0.10	5.89	**EFD30/15/9**
1.34	0.89	19.30	4.60	0.67	1.02	6.70	0.31	0.12	0.11	5.36	**ETD29/16/10**
1.71	1.20	20.90	5.75	1.17	1.66	8.30	0.48	0.23	0.14	6.13	**ETD34/17/11**
2.34	1.73	25.70	6.75	2.17	2.93	13.10	0.88	0.55	0.19	6.97	**ETD39/20/13**
2.79	2.11	29.50	7.15	3.65	4.82	16.90	1.21	1.05	0.22	7.85	**ETD44/22/15**
3.43	2.68	32.70	8.20	5.66	7.25	20.10	1.65	1.74	0.24	8.66	**ETD49/25/16**
4.50	3.64	37.70	9.65	10.19	12.61	25.10	2.42	3.39	0.27	9.80	**ETD54/28/19**
5.19	4.24	42.30	10.03	15.61	19.09	29.70	2.98	5.48	0.29	10.78	**ETD59/31/22**
5.70	4.75	38.00	12.50	37.53	45.01	25.40	3.18	12.54	0.28	14.03	**PM74/59**
8.47	7.32	45.30	16.15	66.58	77.10	32.70	5.28	24.03	0.31	15.87	**PM87/70**
14.18	12.66	60.30	21.00	217.80	243.81	47.70	10.02	86.15	0.35	20.94	**PM114/93**
1.63	1.12	21.90	5.13	0.95	1.37	9.30	0.48	0.20	0.15	5.43	**EC35**
2.15	1.56	25.10	6.23	1.89	2.60	12.50	0.78	0.47	0.18	6.43	**EC41**
3.12	2.42	29.10	8.30	4.35	5.61	16.50	1.37	1.23	0.22	7.65	**EC52**
6.39	5.37	42.80	12.55	14.99	17.84	30.20	3.79	5.29	0.30	9.93	**EC70**

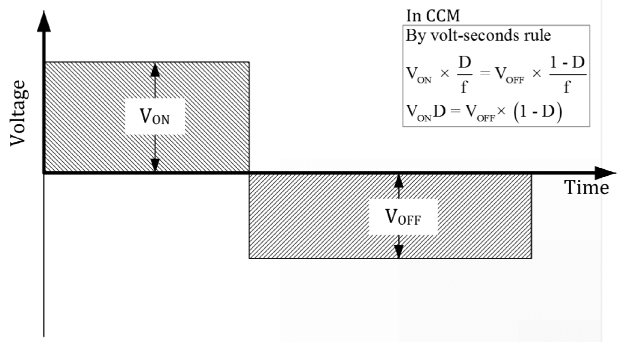
W_{ac} is window area of core; W_{ab} is window area in side bobbin; **Width** is the width of any layer inside bobbin if no margin tape were present; **Height** is the height available for winding copper; $\mathbf{AP_b}$ is the area product of the bobbin; $\mathbf{AP_c}$ is the area product of the core; **Width_tape** is the actual width available for the copper layer with margin tape present; W_{cu} is the net window area available to wind copper (in Primary and Secondary) with margin tape and bobbin considered; $\mathbf{AP_{cu_P}}$ is the area product available for primary winding alone, assuming it is half the total available; $\mathbf{K_{cu_P}}$ is the actual utilization factor for the primary winding (ratio of AP_{cu_P} to AP_c), **MLT** is the mean (or average) length per turn with the bobbin wall thickness and required minimum clearance considered.

Table 12.5 Popular Cores with Area Product, Window Area, Utilization Factor with 6.3-mm Margin Tape

We can see that *the input voltage actually cancels out*. So, in fact, the volt-seconds across the transformer at high-line remain the same at low-line! So does the current swing, and therefore the flux swing. In fact, we can pick *any input voltage* (minimum, maximum, or nominal) and we will get the (same) primary turns if we use

$$N_P = \frac{V_{IN} \times D}{A_e \times f \times \Delta B} \times 10^6$$

RMS of applied voltage across an inductor:



RMS of this waveform is V_{RMS}, where

$$V_{RMS}^2 = V_{ON}^2 D + V_{OFF}^2 (1-D)$$

Substituting for V_{OFF}

$$V_{RMS}^2 = V_{ON}^2 D + \left(\frac{V_{ON}D}{1-D}\right)^2 (1-D)$$

$$= V_{ON}^2 D + \left(\frac{V_{ON}^2 D^2}{1-D}\right)$$

$$= \frac{V_{ON}^2 D(1-D) + V_{ON}^2 D^2}{1-D}$$

$$= \frac{V_{ON}^2 D - V_{ON}^2 D^2 + V_{ON}^2 D^2}{1-D}$$

So

$$V_{RMS} = V_{ON}\sqrt{\frac{D}{1-D}}$$

Used below

or equivalently

$$V_{RMS} = V_{OFF}\sqrt{\frac{1-D}{D}}$$

The induced EMF equation (ignoring signs) is

$$V = N\frac{d\Phi}{dt} = NA\frac{dB}{dt} = L\frac{dI}{dt}$$

So, if a supposedly average voltage V_{AVG} (which happens to be the center of ramp value, say during the ON-time) is applied for Δt (equal to D/f during the ON-time), we get

$$\Delta B = \frac{V_{AVG}\,\Delta t}{NA}$$ (a relationship valid for power supplies)

Note that magnetics vendors usually evaluate their cores by using a symmetrical excitation, and refer to it as B_{AC}, which is the swing in one direction around center zero. We can connect that to power supplies, by talking of the swing around the center of ramp value (see Fig. 7.2). So, we have in general $B_{AC} = \Delta B/2$.

We can therefore also write in general

$$B_{AC} = \frac{V_{AVG}\,\Delta t}{2 \times NA}$$ **Valid**

This can also be expressed either for the ON-time or OFF-time as

$$B_{AC} = \frac{V_{ON}\,D}{2 \times NAf} \text{ or } B_{AC} = \frac{V_{OFF}\,(1-D)}{2 \times NAf}$$ **Valid**

But only if we have an exact square wave excitation (D = 0.5)

$$B_{AC} = \frac{V_{ON}}{4 \times NAf} \text{ or } B_{AC} = \frac{V_{OFF}}{4 \times NAf}$$ **Conditionally valid**

Similarly, when we derive the formula for a sine wave excitation, we get the factor 4.44 in the denominator, but the voltage in the numerator is then the RMS value of the applied sine wave.

Note that even for the exact square wave case (D = 0.5), the RMS is

$$V_{RMS} = V_{ON} \times \sqrt{\frac{D}{1-D}} = V_{ON} \times \sqrt{\frac{0.5}{1-0.5}} = V_{ON}$$

Similarly, we get

$$V_{RMS} = V_{OFF} \times \sqrt{\frac{1-D}{D}} = V_{ON} \times \sqrt{\frac{1-0.5}{0.5}} = V_{OFF}$$

So, only for a perfect square wave excitation (D = 0.5) we get

For a square wave $B_{AC} = \frac{V_{RMS}}{4 \times NAf}$ ⬅ **Commonly seen form (barely applies to power supplies!)**

For a sine wave: $B_{AC} = \frac{V_{RMS}}{4.44 \times NAf}$ ⬅

(B_{AC} is half the total swing, f is the switching/sine frequency)

The term A above is actually A_e (effective area of core)

Figure 12.8 Different equations for finding primary number of turns.

We may still need to use V_{INMIN} for an entirely different reason. We need to confirm that the turns ratio is such that at minimum input, *the duty cycle does not exceed the maximum duty cycle limit of the controller IC*. See the worked example that follows.

The correct equation to use is the more basic form of Faraday's law (Volt-seconds = *NAB*)

$$V_{IN} \times \frac{D}{f_{Hz}} = N_P \times A_{e_{m^2}} \times \Delta B_T$$

$$N_P = \frac{V_{IN} \times D}{f_{Hz} \times A_{e_{m^2}} \times \Delta B_T} = \frac{V_{IN} \times D \times 10^4}{f_{Hz} \times A_{e_{cm^2}} \times \Delta B_T}$$

We will use this in the numerical example.

Worked Example: Flyback and Forward Alternative Design Paths

In a telecommunications application, such as power over ethernet (PoE), we have an input voltage of 36 to 57 V. We want to design a 200-kHz, 12 V @ 11 A (132 W) Forward converter (the controller is limited to a maximum duty cycle of 44 percent as in a typical single-ended type). Select the transformer core, and calculate the primary and secondary number of turns on it. Also select a secondary choke. If the same application and the same control IC are used for a Flyback, what would be the required core size and the number of turns?

Step-by-Step Forward Converter Design

Core Selection

Assume the efficiency will be close to 85 percent. So for an output of 132 W, the input will be 132/0.85 = 155.3 W. We target a flux swing of 0.15 T maximum and a COR current density of 500 A/cm^2 (slightly more aggressive than the 450 A/cm^2 we usually recommend, see pages 298 and 300). So

$$AP_{cu_{P\_cm^4}} = \frac{12.74 \times P_{IN}}{f_{kHz} \times \Delta B_T \times J_{A/cm^2}} = \frac{12.74 \times 155.3}{200 \times 0.15 \times 500} = 0.132 \text{ cm}^4$$

This is the required area product in terms of the primary winding. We expect to use 2-mm margin tape. Therefore, we look at Table 12.3. We see that AP_{cu_P} of 0.13 cm^4 is available from EFD 30/15/9, almost exactly what we need here (0.132 cm^4). That is the selected core.

Primary Turns

We assume the turns ratio will be fixed such that at minimum input the duty cycle is 0.44. So

$$N_P = \frac{V_{IN} \times D \times 10^4}{f_{Hz} \times A_{e_{cm^2}} \times \Delta B_T} = \frac{36 \times 0.44 \times 10^4}{200{,}000 \times 0.69 \times 0.15} = 7.65 \text{ turns}$$

Magnetization Inductance and Peak Magnetization Current

What is the magnetization inductance? The EFD30 with no air gap, made of 3F3 from Ferroxcube, has a datasheet A_L value of 1900 nH/turns^2. So if we use 8 primary turns, we get an inductance of 1900 nH × 8^2 = 121 μH.

Note that an alternative calculation in literature uses

$$L = \frac{\mu\mu_o N^2 \times A_e}{z \times l_e} \qquad \text{(MKS units)}$$

where l_e is the effective length and A_e is the effective area of the core, as defined in Chap. 7. Plugging in our values, we get for primary inductance

$$L = \frac{2000 \times (4\pi \times 10^{-7}) \times 8^2 \times 0.69 \times 10^{-4}}{1 \times 6.8 \times 10^{-2}} = 1.63 \times 10^{-4} \qquad \text{(MKS units)}$$

This is 163 μH.

The difference between the two results is based on the fact that the A_L value provided by the vendor is more practical: It includes the small default air gap since it is not possible to eliminate all air gaps when clamping two separate halves together. So in theory, if there was zero air gap (i.e., an air-gap factor z of 1, see the *A-Z* book), we would get 163 μH. In reality, the magnetization peak current will be higher than expected, because of the minute residual air gap, which has reduced the measured inductance to 121 μH.

So the actual peak magnetization current component in the switch will be a little higher than anticipated (though this will be the same at any input voltage as explained earlier):

$$I_{MAG} = \frac{V_{IN} \times D/f}{L} = \frac{36 \times 0.44/200{,}000}{121 \times 10^{-6}} = 0.655 \text{ A}$$

Turns Ratio

The turns ratio is derived from

$$D = \frac{V_O}{V_{INR}} = \frac{n \times V_O}{V_{IN}}$$

$$n = \frac{D \times V_{IN}}{V_O} = \frac{0.44 \times 36}{12} = 1.32$$

Voltage Ratings

So the maximum reflected input voltage is (see Appendix for voltage stress tables)

$$V_{\text{INRMAX}} = \frac{V_{\text{INMAX}}}{n} = \frac{57}{1.32} = 43.2 \text{ V}$$

The minimum voltage rating of the output diode is

$$V_{\text{D1}} = V_{\text{INRMAX}} + V_O = 43.2 + 12 = 55.2 \text{ V}$$

The minimum voltage rating of the catch diode is

$$V_{\text{D2}} = V_{\text{INRMAX}} = 43.2 \text{ V}$$

If the two diodes are in the same package, we may just get away with a 60-V Schottky with a slight adjustment of the turns ratio to increase headroom.

If this is a single-switch Forward converter, the maximum drain-to-source voltage is twice the input, i.e., 2 × 57 V. So we should look for a 150 V FET. If this is a two-switch Forward (asymmetric half-bridge) the maximum drain-to-source voltage is only 57 V.

Secondary Turns

The number of secondary turns is

$$N_S = \frac{N_P}{n} = \frac{8}{1.32} = 6.06 \approx 6 \text{ turns}$$

Sense Resistor

The peak output current is about 1.2 × 11 A = 13.2 A. This occurs at high-line actually, and the inductor is designed to give a 20 percent peak above average ($r = 0.4$). On the switch, we also need to add the peak magnetization current of 0.655 A. So the sense resistor will have to be set to permit the normal operating current of 13.2 A/n + 0.655 A = (13.2 A/1.333) + 0.655 = 10.6 A. We can set current limiting at about 12 A. So, for example, if the controller IC sense threshold is 200 mV, we need a sense resistor of V/I = 0.2/12 = 0.017 Ω. The RMS current through the sense resistor is about $(11\text{ A}/n) \times \sqrt{D} = (11\text{ A}/1.333) \times \sqrt{0.5} = 5.8$ A. So heating in the sense resistor is about $5.8^2 \times 0.017 = 0.57$ W. For adequate derating we can pick two 33-mΩ/0.5-W resistors in parallel.

Minimum Duty Cycle

We will need this shortly:

$$D_{\text{MIN}} = \frac{V_O}{V_{\text{INRMAX}}} = \frac{n \times V_O}{V_{\text{INMAX}}} = \frac{(8/6) \times 12}{57} = 0.28$$

Choke Inductance and Rating

We have to design this at maximum input because, as in any regular Buck, the maximum peak current occurs at maximum input. At that point we want a total swing ΔI equal to about 40 percent the average value (11 A). This is 20 percent above and 20 percent below the center at I_O.

We need the duty cycle at maximum input from the preceding above step. So, setting a current ripple ratio of 0.4, using the standard Buck equations:

$$L_{\mu\text{H}} = \frac{V_O}{I_O \times r \times f_{\text{Hz}}} \times (1 - D) \times 10^6 = \frac{12}{11 \times 0.4 \times 200{,}000} \times (1 - 0.28) \times 10^6 = 9.82$$

So we pick an inductance of 10 μH. It must have a minimum saturation rating of 12 A.

Overall Loss Estimation in Transformer

From Table 12.3, we see that EFD30 with 2-mm margin tape has a total available area of 0.38 cm^2 for winding copper (both Primary and Secondary). Assuming no Litz wire (no silk covering, etc.) and ignoring wires slipping into adjacent spaces between wires (as in a bundle), the simplest assumption to make here is that 78.5 percent (i.e., π/4) of the physical space is occupied by copper. So the actual area occupied by copper in our case is 0.38 × 0.785 = 0.2983 cm^2.

Let us for now just look at the primary winding, and assume that it occupies half the available window area W_{cu}. So we have 0.2983/2 = 0.15 cm^2 window reserved for primary-side copper. This has $N_P = 8$ turns. Assuming all eight turns are laid out side by side, no wasted space, each turn will have a cross-sectional area of 0.15/8 = 0.01875 cm². The mean length per turn is also given in Table 12.3 for the EFD30 core as 5.89 cm. The length of the entire primary winding is therefore $N_P \times$ MLT = 8 × 5.89 = 47.12 cm. Using the resistivity of copper (17 nΩ · m) we get

$$R = \rho \times \frac{l}{A} = 17.2 \times 10^{-9}\ \Omega\text{m} \times \frac{47.12 \times 10^{-2}\,\text{m}}{0.01875 \times 10^{-4}\,\text{m}^2} = 4.32\ \text{m}\Omega$$

The RMS switch/primary current is about 5.8 A as shown under the sense resistor calculations given earlier. See page 188 too. So the primary side dissipation is

$$P_{\text{cu}_P} = 2 \times 5.8\ \text{A}^2 \times 4.32\ \text{m}\Omega = 0.29\ \text{W}$$

where we have also silently doubled the dissipation *because we have assumed that the AC resistance is, at best, twice the DC resistance*.

With the same current density all through the transformer, we can assume that the dissipation is split equally between the primary and secondary windings. So the total transformer dissipation is finally estimated to be 0.29 × 2, that is, about 0.6 W. But this assumes we can achieve $F_R = 2$. If F_R was closer to 3 (more likely), we would get about 1 W. Further, if copper loss equals core loss, we would then get up to 2-W total transformer dissipation. From Fig. 12.7 we see that an EFD30 transformer (with an effective volume of 4.7 cm³) has a thermal resistance of about 25°C/W. So we could expect a temperature rise of up to 2 W × 25 = 50°C. We would then be approaching the limit of a Class A transformer (55°C allowable rise).

Core Loss and Total Estimated Loss

Assuming we are using 3F3 material from Ferroxcube, the core loss equation is (see Tables 12.6 and 12.7):

$$\text{Core loss} = C \times B^p \times f^d \times V_e \qquad \text{(using system B)}$$

where for 3F3, we have (valid up to 300 kHz):

$$C = 1.3 \times 10^{-16}, \qquad p = 2.5, \qquad d = 2$$

Here B is half the flux swing (in gauss) since vendors use symmetric sine wave excitation for testing, and B refers to the amplitude of the swing (around zero). We get for our current example (EFD30)

$$\text{Core loss} = 1.3 \times 10^{-16} \times 750^{2.5} \times 200{,}000^2 \times 4.7 = 376.5\ \text{mW} \qquad \text{(using system B)}$$

where we have used the core-loss coefficients provided in Table 12.6. We can also derive these as shown in Fig. 12.9. Note that our design point corresponds to (half total flux swing of) 75 mT (750 G), and this corresponds to 100 kW/m³, which is numerically the same as 100 mW/cm³. So we have targeted a conservative core loss of 100 mW/cm³, as most in the industry do (though some claim up to 200 mW/cm³ is OK). Basically, our maximum switching frequency is largely determined by this aspect (core loss).

Constant × $B^{(\text{exponent of } B)}$ × $f^{(\text{exponent of } f)}$ (Core loss per unit volume)							
	Constant	**Exponent of *B***	**Exponent of *f***	***B***	***f***	V_e	**Units**
System A	C_c	C_b	C_f	T	Hz	cm³	W/cm³
	$= \frac{C \times 10^{4\times p}}{10^3}$	$= p$	$= d$				
System B	C	p	d	G	Hz	cm³	mW/cm³
	$= \frac{C_c \times 10^3}{10^{4\times C_b}}$	$= C_b$	$= C_f$				
System C	K_p	n	m	G	Hz	cm³	W/cm³
	$= \frac{C}{10^3}$	$= p$	$= d$				

Table 12.6 The Different Systems in Use for Describing Core Loss (and Their Conversions)

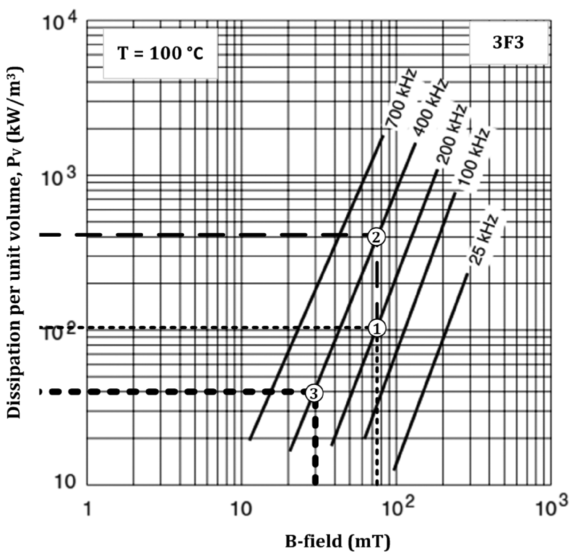
Material (vendor)	Grade	C	p (B^p)	d (f^d)	μ	≈ B_{SAT} (G)	≈ f_{MAX} (MHz)
Powdered iron (Micrometals)	8	4.3E – 10	2.41	1.13	35	12,500	100
	18	6.4E – 10	2.27	1.18	55	10,300	10
	26	7E – 10	2.03	1.36	75	13,800	0.5
	52	9.1E – 10	2.11	1.26	75	14,000	1
Ferrite (Magnetics, Inc.)	F	1.8E – 14	2.57	1.62	3,000	3,000	1.3
	K	2.2E – 18	3.1	2	1,500	3,000	2
	P	2.9E – 17	2.7	2.06	2,500	3,000	1.2
	R	1.1E – 16	2.63	1.98	2,300	3,000	1.5
Ferrite (Ferroxcube)	3C81	6.8E – 14	2.5	1.6	2,700	3,600	0.2
	3F3	1.3E – 16	2.5	2	2,000	3,700	0.5
	3F4	1.4E – 14	2.7	1.5	900	3,500	2
Ferrite (TDK)	PC40	4.5E – 14	2.5	1.55	2,300	3,900	1
	PC50	1.2E – 17	3.1	1.9	1,400	3,800	2
Ferrite (Fair-Rite)	77	1.7E – 12	2.3	1.5	2,000	3,700	1

Note: (a)E-(b) is the same as (a) × $10^{-(b)}$.

Table 12.7 Typical Core-Loss Coefficients of Common Materials (System B)

We see that we may get up to 0.6 to 1-W copper loss and 0.4 W of core loss. They are not equal, and in fact in most modern converters, the assumption of equal copper and core losses is not necessarily true. As mentioned in Chap. 11, it has been reported that a more optimum operating point (minimum core and copper losses combined) is actually defined by

$$\frac{\text{Core loss}}{\text{Copper loss}} = \frac{2}{\text{exponent of } B} \equiv \frac{2}{p} \Rightarrow \frac{2}{2.5} = 0.8 \qquad \text{(for 3F3)}$$



Three points selected in succession as shown: the first two at same B, the last two at same f.

Point 1: 75 mT, 200 kHz, 100 kW/m³
Point 2: 75 mT, 400 kHz, 400 kW/m³
Point 3: 400 kHz, 30 mT, 40 kW/m³

Coordinates

$P_{v1} = 100$, $P_{v2} = 400$, $P_{v3} = 40$

$f_1 = 200$, $f_2 = 400$, $f_3 = 400$

$B_1 = 75$, $B_2 = 75$, $B_3 = 30$

Equations/Calculations

$$\text{exp\_f} = \frac{\ln(P_{v1}/P_{v2})}{\ln(f_1/f_2)} = \frac{\ln(100/400)}{\ln(200/400)} = 2$$

$$\text{exp\_B} = \frac{\ln(P_{v2}/P_{v3})}{\ln(B_2/B_3)} = \frac{\ln(400/40)}{\ln(75/30)} = 2.5$$

Using System B (see accompanying text) (cm³, mW/cm³, Hz, G), and keeping in mind that kW/m³ is numerically equal to W/cm³, we get the constant term C as follows

$$(\text{kW/m}^3) = (\text{constant} \times B^{\text{exp\_B}} \times f^{\text{exp\_f}})$$

$$C = \text{constant} = \frac{\text{kW/m}^3}{B^{\text{exp\_B}} \times f^{\text{exp\_f}}} = \frac{100}{750^{2.5} \times (200000)^2} = 1.6 \times 10^{-16}$$

This is based on eyeballing the above curve.

More exact values are (in System B):

$C = 1.3 \times 10^{-16}$, exponent of B = 2.5, exponent of f = 2

See Pages 106, 107 of *Switching Power Supplies A-Z, 2e*

Figure 12.9 Evaluating the core-loss coefficients from the vendor's core-loss curves, such as for 3F3.

Copper Sizing and Transformer Windings

The width available for the primary winding is 15.7 mm from Table 12.3 (EFD30 with 2-mm margin tape). The skin depth at 200 kHz is (assuming a temperature of 80°C for adjusting the resistivity of copper better)

$$\delta_{\text{mm}} = \frac{66.1[1+0.0042(T-20)]}{\sqrt{f_{\text{Hz}}}} = \frac{66.1[1+0.0042(80-20)]}{\sqrt{200,000}} = 0.185 \text{ mm}$$

When dealing with round wires, we find that just decreasing F_R does not necessarily correlate with lowest F_R because F_R is a ratio. The logic is explained in Fig. 12.10 and leads us to the subdivision strategy shown in Fig. 12.11.

The subdivision strategy is further explained in the *A-Z* book (chapter 3), but as we saw there too, it is not a good idea to start off with an X ($X = h/\delta$) of greater than around 4; otherwise, we have to subdivide too much, never really getting to a low-enough F_R value. And when we do, we will end up with impracticable or nonexistent wire gauges.

To start with, let us assume just a simple winding arrangement of a primary winding followed by a secondary winding. Let us call this a *P-S arrangement*, one primary portion followed by a secondary portion.

P-S Winding Arrangement

Let us start with just a single strand for the Primary and adjust its diameter so it completely fills one layer with eight turns. The layers per portion p for this equals 1. The diameter is width/N_p = 15.7 mm/8 T = 1.963 mm. Note that using the equivalent foil transformation of Dowell, that diameter would give an equivalent foil of thickness $h = 0.886 \times 1.963 = 1.739$ mm (since $\sqrt{\pi/4}$ equals 0.886, see Figs. 10.3 and 11.10). In terms of penetration ratio X, that would be $X = h/\delta = 1.739$ mm/0.185 mm = 9.4. But that is too high a starting value for entering the subdivision process.

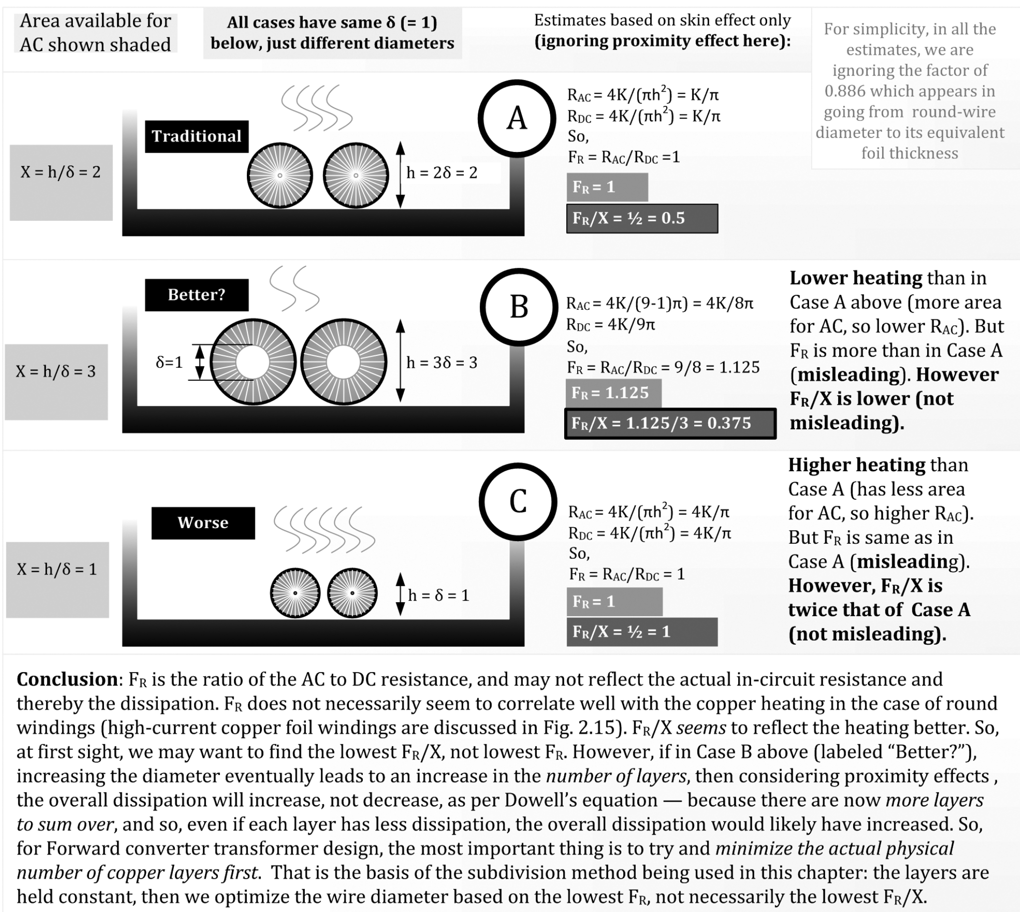


Figure 12.10 F_R/X, not just F_R, correlate better with lowest losses, but the best way is to keep layers unchanged and subdivide strands to achieve lower AC resistance. See Fig. 12.11.

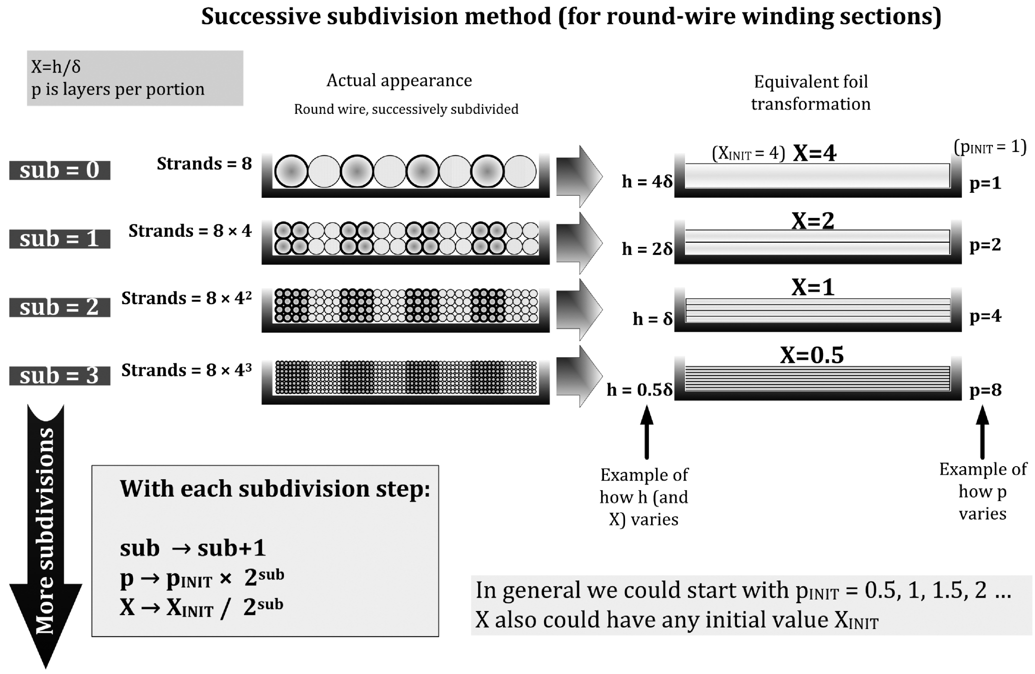


FIGURE 12.11 Subdivision strategy for round wires explained.

So instead, let us start with two paralleled strands of round wire, of diameter half of that, that is, Width/(2 × N_p) = 1.963/2 = 0.98 mm, but still laid out in one layer. So the layers per portion p still equals 1. We say p_{INIT} is 1. Similarly, the starting X value (X_{INIT}) is

$$X_{\text{INIT}} = \frac{0.886 \times \text{Dia}}{\delta} = \frac{0.886 \times 0.98}{0.185} = 4.7$$

Let us go with this for now. From Fig. 12.12 (lower half, that is, $p_{\text{INIT}} = 1$) we see that for $X_{\text{INIT}} = 4.7$ we need six subdivisions to get F_R to fall below 2. Each subdivision splits each strand into four wires, each of half the diameter.

NOTE *In Fig. 12.13, for convenience we have also provided the plots for p_{INIT} equal to 1.5 and 2. These curves are just Dowell's curves plotted out in a way that is useful for the subdivision strategy, and, of course, to which DC bias has been added (which Dowell had not included), and also summed up to 40 Fourier harmonics as shown in Fig. 12.14 (Dowell had just used a high-frequency sine wave in his analysis, as explained).*

After six subdivision steps, we will be left with a wire diameter of $d/(2)^{\text{sub}} = 0.98/2^6 =$ 0.015 mm. In mils this is

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.015}{0.0254} = 0.591 \text{ mils}$$

The nearest AWG is

$$\text{AWG} = 20 \times \log\left(\frac{1000}{\text{mils} \times \pi}\right) = 20 \times \log\left(\frac{1000}{0.591 \times \pi}\right) = 54$$

But that is an impractically thin AWG (if it exists!). The minimum wire gauge usually available is AWG 52, and the thinnest practical value to use for multistrand bundles of magnet wire is AWG 42 to AWG 44).

So we should now start with three paralleled strands instead (in one layer, so $p_{\text{INIT}} = 1$). The starting diameter is 1.963/3 = 0.645 mm. The starting X is

$$X_{\text{INIT}} = \frac{0.886 \times \text{Dia}_{\text{INIT}}}{\delta} = \frac{0.886 \times 0.645}{0.185} = 3.1$$

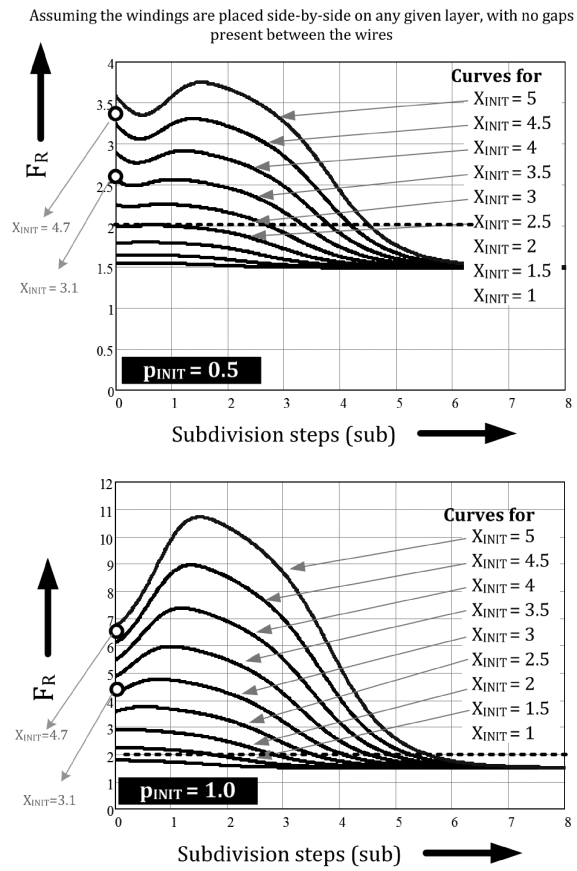


Figure 12.12 Subdivision strategy design curves for p_{INIT} = 0.5 and 1 (see also Fig. 12.16).

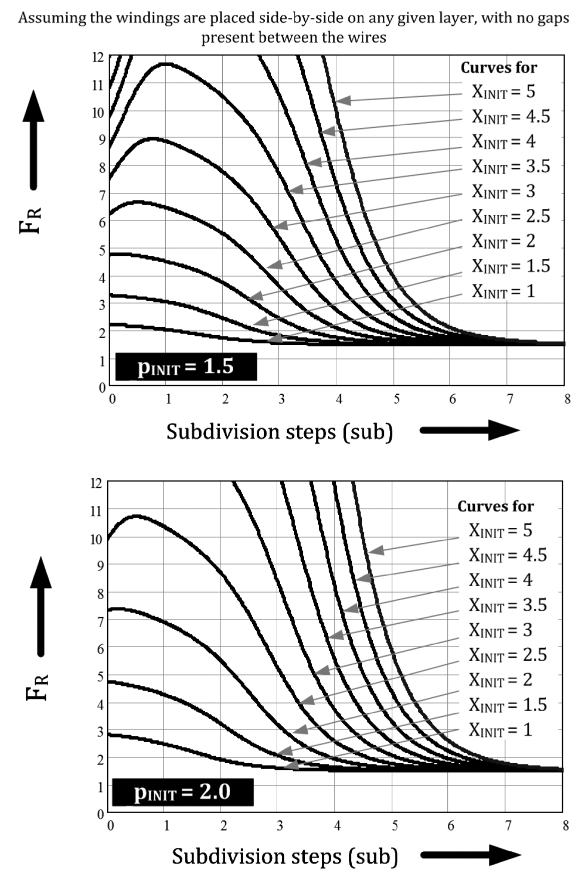


Figure 12.13 Subdivision strategy design curves for p_{INIT} = 1.5 and 2 (see also Fig. 12.16).

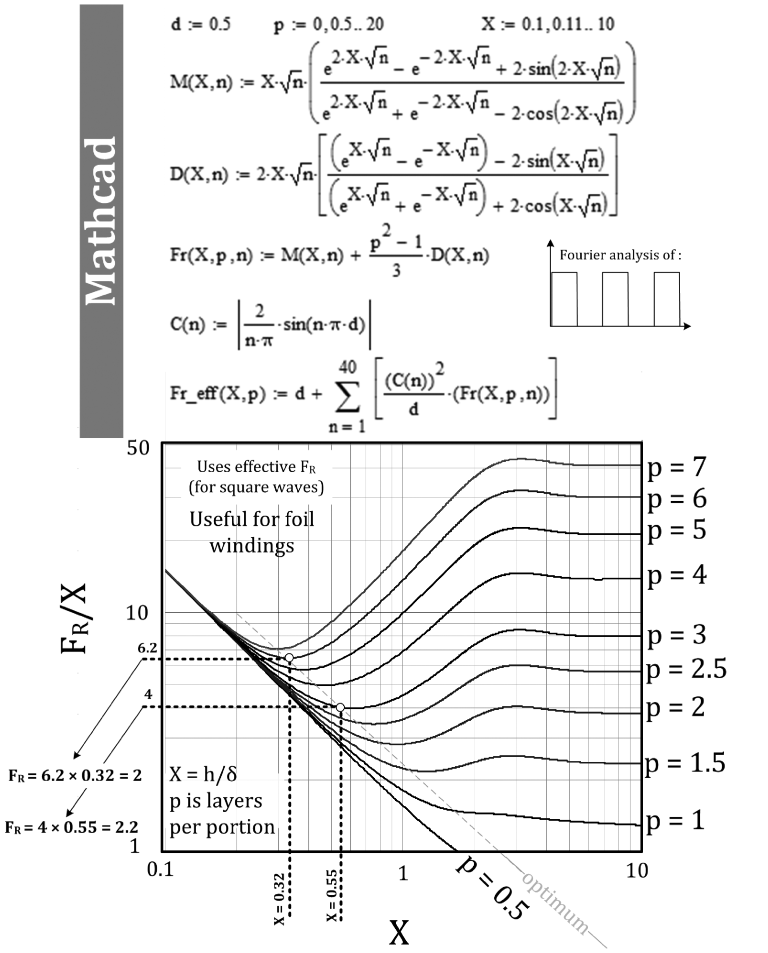


Figure 12.14 Dowell's curves modified for square waves and plotting F_R/X versus X for foil design.

p_{INIT} is still 1. From lower half of Fig. 12.12 we see that for X_{INIT} of 3.1, we need three subdivisions to get F_R to fall close to 3. So we will be left with a diameter of $0.645/2^3 = 0.081$ mm.

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.081}{0.0254} = 3.19 \text{ mils}$$

The nearest AWG is

$$\text{AWG} = 20 \times \log\left(\frac{1000}{\text{mils} \times \pi}\right) = 20 \times \log\left(\frac{1000}{3.19 \times \pi}\right) = 40$$

This is acceptable, though the F_R is around 3, not 2.

In any subdivision step, a single strand becomes four strands. So, the number of strands after splitting each strand *sub* number of times, is $4^{sub} = 4^3 = 64$. So one possible implementation is to use a twisted bundle of magnet wire, consisting of 64 strands of AWG 40. Further, consistent with our starting assumption, three such bundles need to be laid out in parallel (all on one physical layer) to complete eight turns of the primary winding.

For the secondary winding, we look at Fig. 12.15. We see that indeed, optimizing F_R/X is a good idea for foils, especially because, unlike round wires, the layers per portion remain fixed if we increase the foil thickness. To optimize F_R/X, we can consult the lower part of Fig. 12.14. We see that with six layers (turns) per portion, as in our case, we have an optimum $F_R/X = 6.2$ for $X = 0.32$ (corresponding to $F_R = 6.2 \times 0.32 = 2$). So we need a copper foil of thickness $h = X \times \delta = 0.32 \times 0.185$ mm $= 0.059$ mm.

In mils, this is

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.059}{0.0254} = 2.3 \text{ mils}$$

This is the optimum thickness of foil suggested. In general, 1, 1.4, 3, 5, 8, 10, 16, and 22 mils are the more commonly available foil thicknesses, but others can be ordered too.

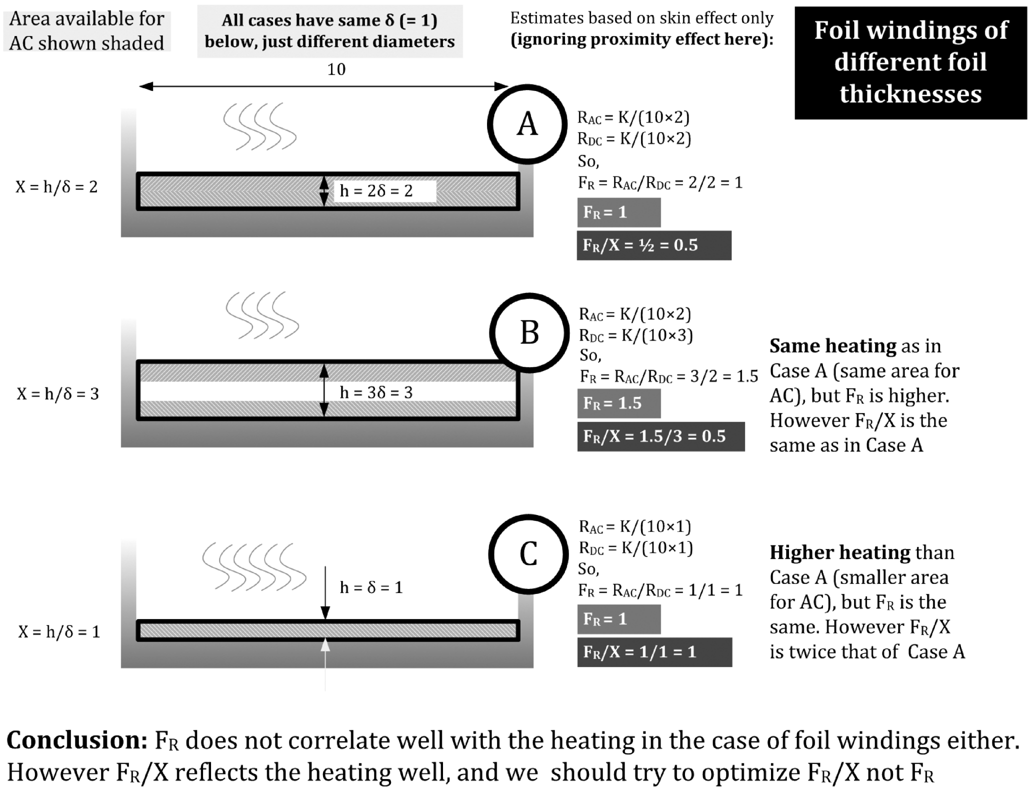


Figure 12.15 Subdivision strategy for foil windings explained.

The width of this foil can be up to 15.7 mm (see "Width_tape" for EFD30 in Table 12.3). So the total copper cross-sectional area is 0.059 mm × 15.7 mm = 0.9263 mm^2. Our secondary-side current is 11 A (COR value), which for $D = 0.5$ is an RMS of $11 \times \sqrt{0.5} = 7.8$ A. If this passes through 0.9263 mm^2, the current density will be 7.8/0.0093 = 838 A/cm^2. At a minimum duty cycle of 0.28 we will get the highest RMS value of $11\text{ A} \times \sqrt{1 - D_{MIN}} = 11 \times \sqrt{0.72} =$ 9.33 A. So the worst-case RMS current density will be 9.33/0.0093 = 1000 A/cm^2. This is well over our target of 250 A/cm^2 RMS current denity, so losses will increase significantly.

So far we have ignored the possibility of interleaving. We have eight turns in the Primary, and we would like to keep that as one physical layer. Instead let us split the Secondary into two *series sections*. We can think of splitting the Secondary into two parallel sections too, but in that case we can get severe EMI due to slight imbalances in the paralleled halves, a sort of "ground loop" phenomena deep inside the transformer, unless we decide to "OR" the paralleled windings though separate output diodes. Also, in paralleled windings, the layers per portion will not decrease; in fact, the number of secondary portions will just double.

So we are now trying a series-split Secondary, sandwiching a single layer of Primary.

S-P-S Winding Arrangement

Let us start with just a single strand for the Primary and adjust its diameter so it completely fills one layer with eight turns. The layers per portion p for this, now equals 1/2, not 1, since each half Primary gets assigned to half the split Secondary on either side. The diameter is width/N_p = 15.7 mm/8T = 1.963 mm. Note that using the equivalent foil transformation of Dowell that diameter would give an equivalent foil of thickness $h = 0.886 \times 1.963 = 1.739$ mm (since $\sqrt{\pi/4}$ equals 0.886). In terms of penetration ratio X, that would be $X = h/\delta = 1.739\text{ mm}/0.185\text{ mm} = 9.4$. But that is too high a starting value for entering the subdivision process.

So instead, let us start with two paralleled strands with round wire, with diameter half of that, i.e., width/$(2 \times N_p)$ = 1.963/2 = 0.98 mm, but still laid out in one layer. So the layers per portion p still equals 1/2. We say p_{INIT} is 1/2. Similarly, the starting X value (X_{INIT}) is

$$X_{INIT} = \frac{0.886 \times \text{Dia}}{\delta} = \frac{0.886 \times 0.98}{0.185} = 4.7$$

Let us go with this for now. From Fig. 12.12 (upper half, that is, $p_{INIT} = 1/2$) we see that we need three subdivisions to get F_R to fall below 3. That could be OK, but we also look at the following case.

Suppose we start with three paralleled strands instead (in one layer, so $p_{INIT} = 1/2$). The starting diameter is 1.963/3 = 0.645 mm. The starting X is

$$X_{INIT} = \frac{0.886 \times Dia_{INIT}}{\delta} = \frac{0.886 \times 0.645}{0.185} = 3.1$$

p_{INIT} is still 1/2. From Fig. 12.12 (upper half), we see that for X_{INIT} of 3.1, we *do not need any subdivisions* to get F_R to fall close to 3. In fact we are starting at $F_R = 2.5$ already. So no further subdivision is required.

In mils the strand diameter is

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.645}{0.0254} = 25.4 \text{ mils}$$

The nearest AWG (for strand of diameter) is

$$\text{AWG} = 20 \times \log\left(\frac{1000}{\text{mils} \times \pi}\right) = 20 \times \log\left(\frac{1000}{25.4 \times \pi}\right) = 22$$

So *the Primary consists of three paralleled strands of AWG 22*.

Let us check whether we can accommodate this. In all, we will have 8 × 3 = 24 strands side by side in one layer. So it will occupy 0.645 × 24 = 15.5 mm. We know from Table 12.3 that we have 15.7 mm available. So this is acceptable.

NOTE *Coming to the Secondary, we could consider trying to avoid a foil winding if possible (for reasons of cost). We can then proceed as we did for the Primary. But keep in mind that p_{INIT} is no longer equal to 3 for the Secondary if we use round wires instead of a foil.*

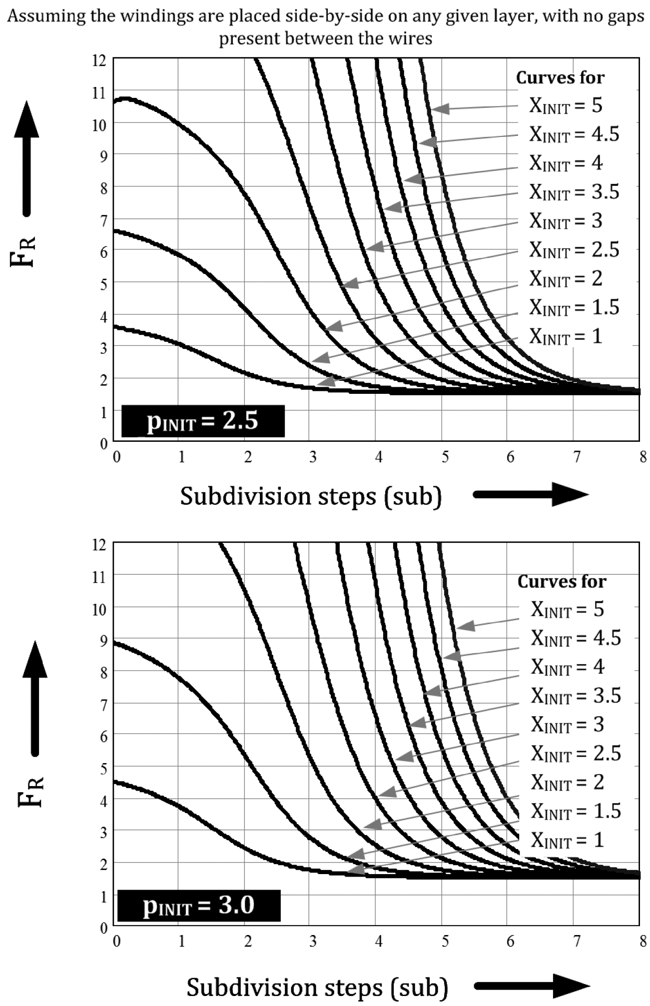


FIGURE 12.16 Subdivision strategy design curves for $p_{INIT} = 2.5$ and 3.

Let us return to three turns of foil on each side of the sandwiched Primary. We look at Fig. 12.14, along the $p = 3$ curve this time. We see that we have an optimum $F_R/X = 4$ for $X = 0.55$ (corresponding to $F_R = 4 \times 0.55 = 2.2$). So we need a copper foil of thickness $h = X \times \delta = 0.55 \times 0.185$ mm = 0.102 mm.

In mils, this is

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.102}{0.0254} = 4.0 \text{ mils}$$

We look for a 4-or 5-mil-thick copper foil.

The width of this foil can be up to 15.7 mm (see "Width_tape" for EFD30 in Table 12.3). So the total copper cross-sectional area is 0.102 mm × 15.7 mm = 1.6 mm². Our secondary-side current is 11 A (COR value), which for $D = 0.5$ is an RMS of $11 \times \sqrt{0.5} = 7.8$ A. If this passes through 1.6 mm², the RMS current density will be 7.8/0.016 = 487 A/cm². This is within the usual industry target of 500 A/cm² at least (see page 299). But the losses are going to be higher than we expected, based on $F_R = 2$. Our initial target was 500 A_{COR}/cm² on page 315.

Keep in mind that in a foil winding, as explained in Fig. 12.15, even if we increase the foil thickness, since the skin and proximity effects restrict the effective cross-sectional area actually passing high-frequency current, that area remains fixed if we increase foil thickness beyond a certain point. So AC resistance will not improve. In reality because of *excess copper* exposed to higher proximity effects (higher X), we could end up worsening the situation. So how do we improve the current density for foils? The only possibility is by choosing cores with *long* profiles. For this we should also consider EER/ER and EERL/ERL cores in particular.

Input Capacitor Selection

In AC-DC applications, this is a wide topic involving holdup time, power factor correction, etc. Here we are dealing only with a DC-DC converter for telecommunications applications. So, the primary (dominant) selection criterion is the RMS current. We first select a bulk capacitor, preferably aluminum electrolytic, for cost reasons. The input RMS of a Buck (at $D \approx 0.5$) is $I_O/2$ (see page 712 of the *A-Z* book). In a Forward, this is reflected to the Primary through the turns ratio, as $I_O/2n$, i.e., $I_{OR}/2$. So, ignoring the small magnetizing current, we need a capacitor with an RMS rating of $I_O/n = 11 \text{ A}/(2 \times 1.33) = 4.1$ A. We tentatively pick UVY1J102MHD from Nichicon. This is a 105°C, 1000-h, 1000-µF, 63-V capacitor with a stated RMS current capability of 0.93 A. At high frequencies we can apply the typical frequency multiplier of $\sqrt{2} = 1.414$ (Nichicon actually allows 1.5). So its high-frequency RMS rating is, conservatively, $0.93 \text{ A} \times \sqrt{2} = 1.32$ A. If we parallel three of these, we get $1.32 \times 3 = 4$ A, which is what we need. At a maximum ambient of 45°C, we can add a worst-case additional 10°C rise from local heating (hot components, enclosure, etc.), so we estimate the surface of the capacitor can be at 55°C. We know that every 10°C below the upper category temperature (105°C here), we get a doubling of life (provided we do not exceed its datasheet value of ripple current rating at room temperature). So we expect life expectancy to be $1000 \text{ h} \times 2^{(105-55)/10} = 1000 \text{ h} \times 2^5 = 320$ kh, which is about 3.7 years when operated 24 h a day. If we need more life, we need to pick a 2000 h at 105°C capacitor.

The ESR of each capacitor can be determined from its stated tangent of loss angle, $\tan\delta = 0.1$ (here δ is not the skin depth but the loss angle, expressed by this vendor, and most other vendors too, at 120 Hz). The relationship is

$$\tan\delta = \frac{\text{ESR}}{X_C} = \text{ESR} \times 2 \times \pi \times f \times C$$

Solving

$$\tan\delta = \frac{\text{ESR}}{X_C} = \text{ESR} \times 2 \times \pi \times f \times C$$
$$\text{ESR} = \frac{\tan\delta}{2 \times \pi \times f \times C} = \frac{0.1}{2 \times \pi \times 120 \times 1000} \times 10^6 = 0.133\ \Omega$$

Note that this is the ESR at 120 Hz. We can assume the ESR of an aluminum electrolytic capacitor gets better by a factor of 2 at high frequencies (this is the origin of the frequency multiplier $\sqrt{2}$). In addition, we have three capacitors in parallel. So the net high-frequency ESR is 133 mΩ/6 = 22 mΩ.

We therefore have a net capacitance of 3000 μF with an ESR of 22 mΩ. This is acceptable, but where space is of concern, we would like to reduce the bulk capacitance, by paralleling several ceramic capacitors. The technique to do this will now be explained.

Paralleling Ceramic and Electrolytic Capacitors at the Input

Looking at Fig. 12.17 we see that there will be two contributions to the ESR: one from the capacitance and one from the ESR. In an electrolytic capacitor, generally the first contribution is negligible compared to the second. We have a reflected load current of I_{OR} = 8.2 A (I_{OR} for a Forward, I_O for a Buck), split in three capacitors, each therefore supporting 8.2 A/3 = 2.73 A load current. The high-frequency ESR of each is 22 mΩ × 3 = 66 mΩ. With a current through each capacitor of peak-to-peak value $I_{OR}(1 + r/2)/3$ (as per Fig. 12.18, but with three capacitors in parallel sharing I_{OR}), we will get a peak-to-peak ripple of [8.2 A × (1 + (0.4/2))/3] × ESR = 216.5 mV. In fact each identical capacitor produces this very ripple voltage, and these voltages are all in parallel so they do not pass current between one another (ideally). When we use several ceramic capacitors in parallel to replace one or two of the aluminum electrolytic capacitors ("Elkos"), the first thing we have to do to avoid upsetting this "apple cart" is to ensure they produce a ripple smaller than 216.5 mV. If the ripple they produce is more than this, the incoming current will prefer to shift more current through the remaining Elko(s). That will pass excess RMS current through them. But if less ripple is created in parallel, the ceramic capacitors will start taking up more and more of the current, and we may even be able to reduce the size of the last remaining Elko(s). So, let us target 150 mV for the ceramic capacitor combination.

Suppose we have three ceramic capacitors in parallel, their ESR will be typically 20 mΩ/3 = 7 mΩ, since each is about 20 mΩ. This is a rather small contribution to the voltage ripple. The main contribution to the ripple in this case will come from the fact that their net capacitance is not as large as for aluminum capacitors.

We have to now use the capacitance-based equation in Fig. 12.17 for the ceramic capacitors. *We are trying to get rid of two of the three Elkos*. The replacement ceramic combination must support at least 2 × 8.2 A/3 = 4.47 A (shared) load current. Setting a peak-to-peak ripple target of 150 mV, we get

$$150\text{ mV} = \frac{I_{OR} \times D \times (1-D)}{f \times C_{IN}} = \frac{4.47\text{ A} \times 0.5 \times (1-0.5)}{200{,}000 \times C_{IN}}$$

Solving for capacitance,

$$C_{IN} = \frac{4.47 \times 0.5 \times (1-0.5)}{200{,}000 \times 150 \times 10^{-3}} = 3.7 \times 10^{-5}\text{ F}$$

This is 37 μF (net value). To keep the ESR contribution low (well below 33 mΩ, i.e., whatever they are replacing: two 66-mΩ capacitors in parallel here), we should try two or three ceramic capacitors in parallel. Also, knowing that the actual value of a ceramic capacitor in any application may only be 60 percent the printed value in its datasheet, we should aim for almost twice the calculated value we obtained for choosing the printed value.

So finally, a possible solution is: One 1000-μF, 63-V aluminum Elko, in parallel with two 33-μF, 100-V ceramic capacitors. (The two paralleled 33-μF ceramic capacitors will give an in-circuit net value of about 37 μF.)

Output Capacitor Selection

The output is a simple Buck stage. We are trying to choose ceramic capacitors here. There are three main criteria we need to satisfy. This is explained in Fig. 12.18.

1. Maximum peak-to-peak output ripple to be within 1 percent (i.e., ±0.5 percent) of output rail, i.e., $V_{O\_RIPPLE\_MAX}$ = 12 V/100 = 0.12 V.
2. Maximum acceptable droop during a sudden increase in load: ΔV_{DROOP} = 0.5 V.
3. Maximum acceptable overshoot during a sudden decrease in load: $\Delta V_{OVERSHOOT}$ = 0.5 V.

We have minimum output capacitances based on criteria 1 to 3. For criteria 1,

$$C_{O\_MIN\_1} = \frac{r \times I_O}{8 \times f \times V_{O\_RIPPLE\_MAX}} = \frac{0.4 \times 11}{8 \times 200 \times 10^3 \times 0.12} = 2.3 \times 10^{-5}\text{ F}$$

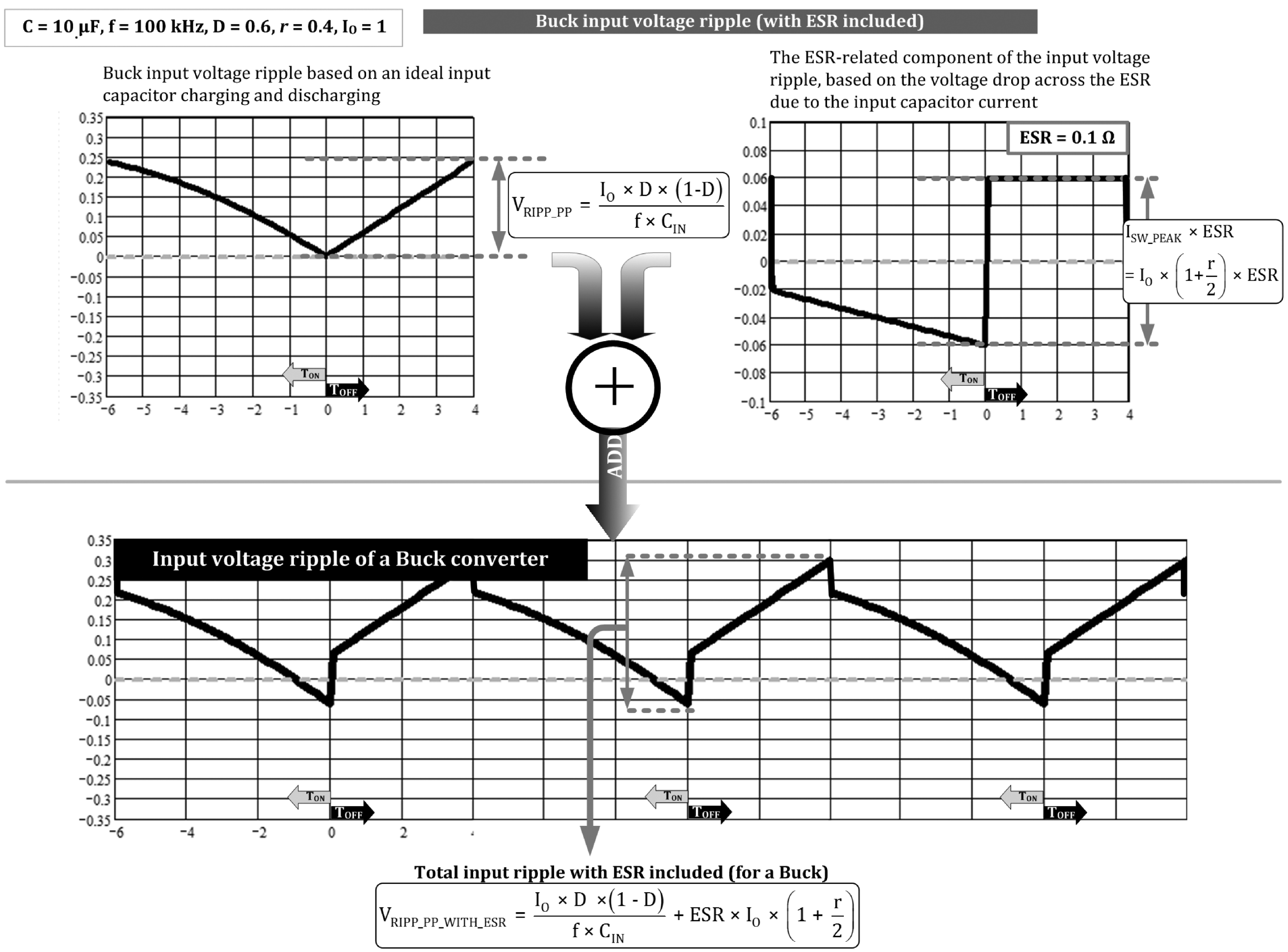


Figure 12.17 Input capacitor waveforms of a Buck.

Maximum ESR: based on maximum output ripple

Ignoring ESR and ESL, purely based on capacitance, the maximum allowed output ripple determines a minimum output capacitance.

$$C_O = \frac{r \times I_O}{8 \times f \times V_{RIPPLE}}$$ (see Fig. 13.5 in *Switching Power Supplies A-Z, 2e*)

$$C_O \geq \frac{r \times I_O}{8 \times f \times V_{RIPPLE\_MAX}}$$

Including ESR, but assuming C is large and ESL is negligible. The maximum allowed voltage ripple determines a maximum ESR

$$V_{RIPPLE} = ESR \times I_O \times r$$

$$ESR \leq \frac{V_{RIPPLE\_MAX}}{I_O \times r}$$

Minimum capacitance: based on maximum droop

Typically, with a well-designed loop, it takes about three switching cycles for the loop to react and start correcting the output to meet a sudden load demand. During that time we do not want the output capacitor to fall more than a certain value V_{droop}. Thus, using I = C dV/dt, we get

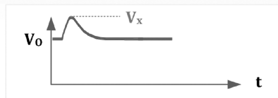
$$I = C\frac{\Delta V}{\Delta t} \quad \Rightarrow \quad C \geq \frac{I \times \Delta t}{\Delta V} = \frac{I \times 3T}{\Delta V_{droop}} = \frac{I \times 3}{\Delta V_{droop} \times f}$$

Here the droop is actually related to the extra load demand, since the normal load requirement is being met every cycle without any droop. So the current here is actually the load increase.

$$C_O \geq \frac{3 \times \Delta I_O}{\Delta V_{droop} \times f}$$

Minimum capacitance: based on maximum overshoot

There is another criterion. In case of a sudden release of load demand, say from maximum load I_o to zero, the inductor energy will all get dumped into the output capacitor. If we do not want too much of an overshoot (say, to restrict the output to some value V_x):



$$\frac{1}{2} \times C\left(V_X^2 - V_O^2\right) = \frac{1}{2} \times L\left(I_O^2\right) \quad \Rightarrow \quad C \geq \frac{L\left(I_O^2\right)}{\left(V_X + V_O\right) \times \left(V_X - V_O\right)} \approx \frac{L\left(I_O^2\right)}{\left(2V_O\right) \times \left(\Delta V_{overshoot}\right)}$$

$$C_O \geq \frac{L\left(I_O^2\right)}{\left(2V_O\right) \times \left(\Delta V_{overshoot}\right)}$$

(where we have used the approximation $V_x + V_o \approx 2 \times V_o$. Also, $V_x - V_o = \Delta V_{overshoot}$)

Figure 12.18 Output capacitor selection criteria for Buck (and Forward) converters.

That is 23 μF. Now based on the second criterion,

$$C_{O\_MIN\_2} = \frac{3 \times (I_O/2)}{\Delta V_{DROOP} \times f} = \frac{3 \times (11/2)}{0.5 \times 10^6} = 3.3 \times 10^{-5}$$

That is 33 μF. Now based on the third criterion,

$$C_{O\_MIN\_3} = \frac{L \times I_O^2}{2 \times V_O \times \Delta V_{OVERSHOOT}} = \frac{10 \times 10^{-6} \times 11^2}{2 \times 12 \times 0.5} = 1.00 \times 10^{-4}$$

This is 100 μF. We need to account for tolerances, etc. Let us therefore pick an output ceramic capacitance of 120 μF/16 V. This will satisfy all the criteria. This may have a large ESR, so we may want to pick paralleled capacitors.

We should double-check that the ESR of the selected capacitor is small enough. The ESR should be less than

$$ESR_{Co\_MAX} = \frac{V_{O\_RIPPLE\_MAX}}{I_O \times r} = \frac{0.12}{11 \times 0.4} = 0.027\ \Omega \qquad \text{(i.e., 27 m}\Omega\text{)}$$

Most small-capacitance ceramic capacitors will have no trouble complying with this. But in our case we may prefer to pick two 56-μF or 68-μF/16-V ceramic caps in parallel, to keep ESR down.

Note that as explained in Fig. 12.19, the contributions to output voltage ripple are not *in phase*. So, for example, if the ESR-based ripple is 120 mV, and the capacitance-based ripple is 120 mV, the total ripple is still around 120 mV, not 240 mV. This is unlike the input.

Step-by-Step Flyback Converter Design

Here the requirements are the same as for the preceding Forward converter. This exercise will give us insight into how a Flyback compares with a Forward, in terms of design methodology and component selection, especially at these high power levels.

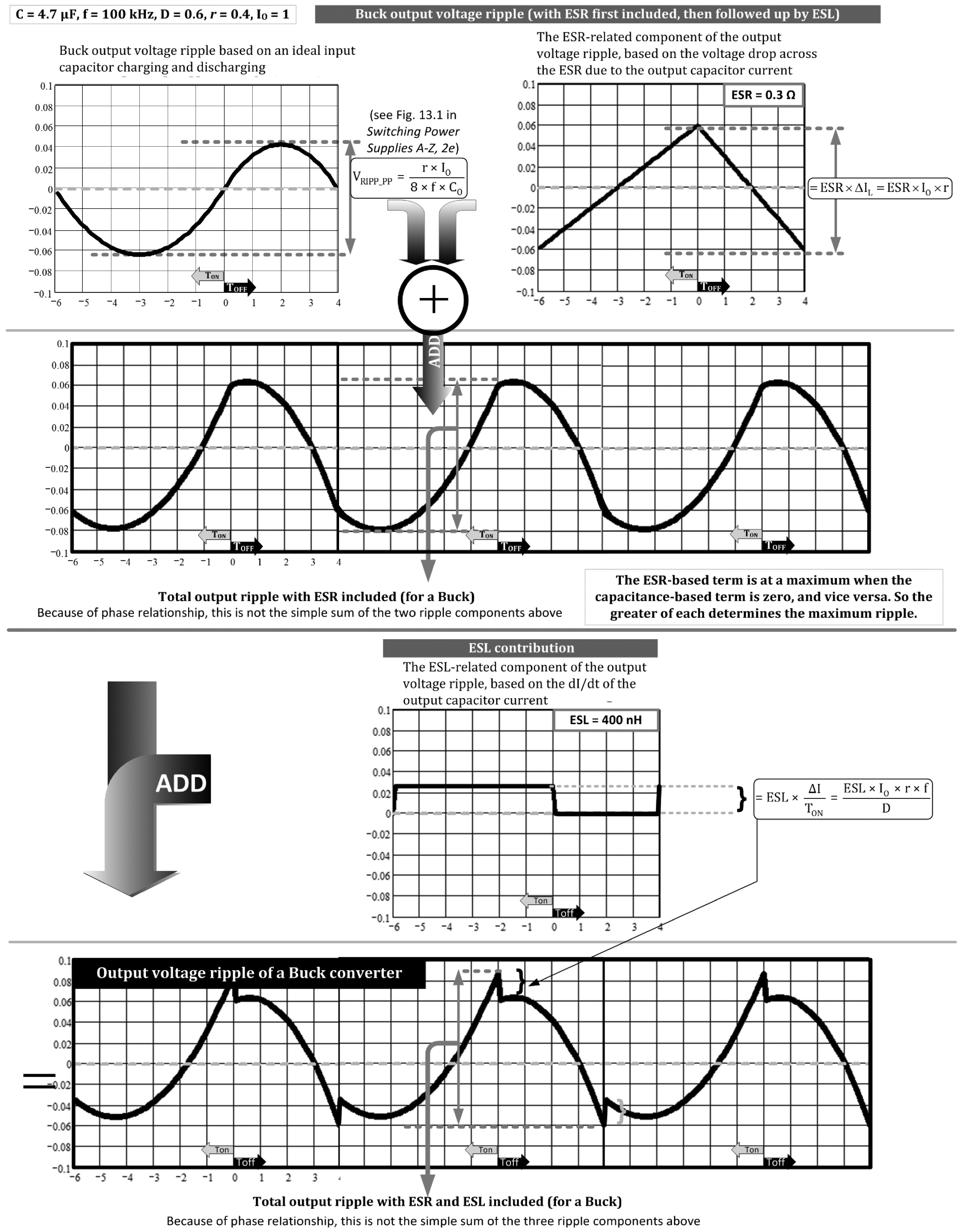


FIGURE 12.19 Output capacitor waveforms of a Buck.

Choosing V_{OR}

Once again, assume the efficiency will be close to 85 percent. So for an output of 132 W, the input will be 132/0.85 = 155.3 W. This is to compare apples to apples, though a Flyback will have much lower efficiency at these power levels, largely due to the huge pulsating current into the output capacitors and leakage inductance dissipation.

We need to set the reflected output voltage (the effective output rail as seen by the primary side). This is also based on the maximum duty cycle limit condition at low-line. We have (see the DC transfer function equation of a Buck-boost in the Appendix)

$$V_{OR} = V_{INMIN} \times \frac{\eta_{VINMIN} \times D_{MAX}}{1 - D_{MAX}} = 36 \times \frac{0.85 \times 0.44}{1 - 0.44} = 24.04 \text{ V}$$

Turns Ratio

Therefore, the turns ratio must be

$$n = \frac{V_{OR}}{V_O} = \frac{24.04}{12} = 2$$

Core Selection

$$V_{e_{cm^3}} = \frac{31.4 \times P_{IN} \times \mu}{z \times f_{MHz} \times B_{SAT_G}^2}\left[r \times \left(\frac{2}{r}+1\right)^2\right] = \frac{31.4 \times 155.3 \times 2000}{10 \times 0.2 \times 3000^2}\left[0.4 \times \left(\frac{2}{0.4}+1\right)^2\right] = 7.8$$

Here we have used the equation derived in *Switching Power Supplies A-Z*, 2d ed (page 225). We have set relative permeability to 2000, maximum saturation flux density to 3000 G (0.3 T), air-gap factor z to 10, and current ripple ratio to 0.4. We need a core volume of 7.8 cm^3. Looking at Table 12.1 we see that the EFD30 we selected for the Forward converter has a volume of 4.7 cm^3. We need almost twice that here. From Table 12.1 we see that a close fit is ETD34 with a volume of 7.64 cm^3 and an effective area of 0.97 cm^2.

Primary Turns

As derived in the *A-Z* book (page 236),

$$N_P = \left(1+\frac{2}{r}\right) \times \frac{V_{INMIN} \times D_{MAX}}{2 \times B_{SAT_T} \times A_{e_{m^2}} \times f_{Hz}}$$

$$= \left(1+\frac{2}{0.4}\right) \times \frac{36 \times 0.44}{2 \times 0.3 \times 0.97 \times 10^{-4} \times 200{,}000} = 8.2 \approx 8 \text{ turns}$$

Secondary Turns

$$N_S = \frac{N_P}{n} = \frac{8}{2} = 4 \text{ turns}$$

Note that the turns ratio is 8/4 = 2, as compared to 1.33 for the Forward converter. This helps pick lower voltage components on the secondary side since the reflected input voltage is lower.

Primary Inductance

From the Appendix, and as derived in *A-Z* book (see page 139),

$$L_{P\_\mu H} = \frac{V_{OR}}{I_{OR} \times r \times f_{Hz}} \times (1-D_{MAX})^2$$

$$= \frac{24.04}{(11/2) \times 0.4 \times 200{,}000} \times (1-0.44)^2 = 1.714 \times 10^{-5}$$

So we need a primary inductance of 17.14 μH.

Zener Clamp

For good efficiency, the zener clamp voltage must be greater than 1.4 times the reflected output voltage. So the minimum recommended clamp voltage is 1.4 × V_{OR} = 1.4 × 24.04 = 33.7 V. But see the subsequent final choice below (i.e., 58 V).

Voltage Ratings

This clamp would require a minimum FET rating of 57 V + 33.7 V= 90.7 V. But if we pick a 100-V FET, there is very little headroom (margin). So we actually pick a 150-V FET instead, and also can then use a 58-V zener clamp. That will give us a maximum drain-to-source voltage of 57 V + 58 V = 115 V. That gives us a good derating margin of 115 V/150 V = 0.77 (i.e., headroom of 23 percent).

We assume the typical leakage is 1 percent of the primary-side inductance, L_{lk} = 171 nH. The peak primary current at low-line is

$$I_{PK\_PRI} = \frac{I_{OR}}{1-D_{MAX}} \times \left(1+\frac{r}{2}\right)$$

$$= \frac{11/2}{1-0.44} \times \left(1+\frac{0.4}{2}\right) = 11.8$$

The zener clamp dissipation will be

$$P_Z = \frac{1}{2} \times L_{lk} \times I_{PK}^2 \times f \times \frac{V_Z}{V_Z - V_{OR}}$$
$$= \frac{1}{2} \times 171 \times 10^{-9} \times 11.8^2 \times 200{,}000 \times \frac{58}{58-24} = 4.1 \text{ W}$$

In reality, if we do bench measurements, we will likely find that the peak current that actually freewheels into the clamp is less. It can be anywhere between 0.7 and 0.9 of the expected peak current, on account of the parasitic capacitances in the transformer, especially in the case of multilayered primary windings. If that happens, the zener clamp dissipation can even be $0.7^2 = 0.5$ (half) what we are expecting from the preceding equation.

The reflected input voltage is

$$V_{INRMAX} = \frac{V_{INMAX}}{n} = \frac{57}{2} = 28.5 \text{ V}$$

The minimum output diode voltage is therefore

$$V_{D1} = V_{INRMAX} + V_O = 28.5 + 12 = 40.5 \text{ V}$$

The currents are very high, so certainly a single diode will not serve the purpose. We will need paralleled diodes, and very likely low-R_{DS} synchronous FET rectification.

Input Capacitor Selection

The equation for the RMS current of an input capacitor of a Buck is

$$I_{IN\_RMS} = I_O \sqrt{D\left(1 - D + \frac{r^2}{12}\right)}$$

so for $D = 0.5$ and r small, we get

$$I_{IN\_RMS} \approx \frac{I_O}{2}$$

For a Forward converter, we just replaced I_O with $I_{OR} = I_O/n$. Similarly, for a Buck-boost we have

$$I_{IN\_RMS} = \frac{I_O}{1-D} \sqrt{D\left(1 - D + \frac{r^2}{12}\right)},$$

so for $D = 0.5$ and r small, we get

$$I_{IN\_RMS} \approx I_O$$

For a Flyback converter, we just replace I_O with $I_{OR} = I_O/n$.

We see that for duty cycles close to 50 percent, the Flyback has twice the input capacitor RMS, compared to a Forward. Another way of looking at this is the Flyback of x W, is equivalent to a Forward of $2x$ W. So, we can repeat the calculations we did for a Forward converter, but mentally visualizing it as a 12 V @ 22 A converter now. All our calculated input capacitances will double. We can similarly trade-off ceramics for some of the Elkos. We conclude that a possible solution is two 1000-µF/63-V aluminum Elkos, in parallel with four 33-µF/100-V ceramic capacitors.

Output Capacitor Selection

In a Flyback and Buck-boost we have a pulsating current into the output capacitors too. It is not *smoothened* by an inductor along the way, as in a Buck or Forward. So the dominant criterion is simply based on the need to be able to absorb this high RMS, without overheating. Any further reduction in output voltage ripple, if necessary, usually comes from a small post-LC filter placed after the initial output capacitors just after the output diode, which are the ones that really take the entire brunt of the output diode current.

The calculation is actually very similar that for to the input capacitor. The equation for the output capacitor RMS of a Buck-boost is

$$I_{O\_RMS} = I_O \sqrt{\frac{D + r^2/12}{1-D}}$$

so for $D = 0.5$ and r small, we get

$$I_{O\_RMS} \approx I_O$$

We need a net capacitor RMS rating of 11 A! We are getting no help from the turns ratio here, as in the case of input capacitors with pulsating currents. Looking at the catalog, we see that rather than use traditional aluminum electrolytic capacitors with huge capacitance values, since we only need capacitors for less than 25 V, a very good candidate is APXC160AR-A820MH70G from Chemicon. It is a 82-μF/16-V *conductive polymer aluminum solid capacitor* with a RMS rating of 2.83 A_{RMS} (at high frequencies), due to its extremely low ESR of 25 mΩ. We need four of these in parallel for a 2.83 × 3 = 11.32 A_{RMS} rating.

Copper Windings

The skin depth at 200 kHz is (assuming a temperature of 80°C for setting the resistivity of copper more accurately)

$$\delta_{mm} = \frac{66.1[1+0.0042(T-20)]}{\sqrt{f_{Hz}}} = \frac{66.1[1+0.0042(80-20)]}{\sqrt{200,000}} = 0.185 \text{ mm}$$

We choose a round wire of diameter 2δ. So we look for a wire of cross-sectional area 0.185 × 2 = 0.37 mm. In mils the strand diameter is

$$\text{mils} = \frac{\text{mm}}{0.0254} \Rightarrow \frac{0.37}{0.0254} = 14.6 \text{ mils}$$

The nearest AWG (for strand of diameter) is

$$\text{AWG} = 20 \times \log\left(\frac{1000}{\text{mils} \times \pi}\right) = 20 \times \log\left(\frac{1000}{14.6 \times \pi}\right) = 27$$

We choose AWG 27. Its cross-sectional area is

$$\text{Area}_{\text{AWG}} = \frac{\pi \times D^2}{4} = \frac{\pi \times 0.37^2}{4} = 0.11 \text{ mm}^2$$

At a target current density of 250 A_{RMS}/cm^2, we can pass 250 × 0.11/100 = 0.275 A_{RMS}. However, at low line, our COR current is $I_{OR}/(1 - D_{MAX})$ = 11 A/[2 × (1 − 0.44)] = 9.82 A. Its RMS value is 9.82 A × $\sqrt{D}$ = 9.82 A × $\sqrt{0.44}$ = 6.5 A_{RMS}. So the number of strands we need for the Primary are 9.82/0.275 = 36 strands. If we double the current density target to 500 A_{RMS}/cm^2, we can go in for 18 strands. Also, as explained in Fig. 12.10, in the case of round wires, if we go to higher diameters than 2δ, we do get an improvement in AC resistance, even though F_R worsens. So we can in fact judiciously go in for thicker wire gauges (lesser number of strands), to fill in each layer fully, if required, and thus get a better build.

On the secondary side, we use the same wire gauge. At $D = 0.44$, the RMS of the secondary current is $I_O/\sqrt{(1-D)}$ = 11 A/$\sqrt{(1-0.44)}$ = 14.7 A_{RMS}. So if we are targeting 250 A/cm², we know that AWG 27 is only capable of 0.275 A_{RMS}. In that case the number of strands required is 14.7/0.275 = 53 strands. If we decided we can double the current density, we can go in for 26 strands for the Secondary.

Note that we have not touched upon the topic of proximity effects in the Flyback, since most agree it is an extremely difficult problem to tackle through closed-form equations. Instead we are just sticking to current density targets. This is discussed next.

Keep in mind that split and sandwiched windings help here too, but mainly to reduce leakage inductance and reduce zener clamp dissipation. Otherwise our underlying assumption of leakage inductance being just 1 percent of the primary inductance won't be true.

Industrywide Current Density Targets in Flyback Converters

In the *A-Z* book, we suggested 400 cmil/A as a recommended current density for the Flyback. See its nomogram and contained explanation on page 145 of the *A-Z* book. That was based on the COR value. To make that clearer here, as per our current terminology, we prefer to write it as 400 cmil/A_{COR}.

Assuming $D \approx 0.5$, we have $\sqrt{D} = 0.707$, so the conversions are

$$\frac{400 \text{ cmil}}{A_{COR}} \equiv \frac{600}{0.707} = \frac{565 \text{ cmil}}{A_{RMS}}$$

or

$$\frac{197{,}353}{400} = \frac{493\ A_{COR}}{cm^2} \quad \text{(in terms of COR current)}$$

or

$$\frac{197{,}353}{565} = \frac{350\ A_{RMS}}{cm^2} \quad \text{(in terms of RMS current)}$$

In other words, we were recommending somewhere between 250 (conservative) to 500 A_{RMS}/cm^2 (overly aggressive). But a lot depends on core losses too, because we should remember, the flux swing in a typical Flyback is always fixed at around 3000 G, not 1500 G as in a Forward converter. So core losses can be four times that of a Forward converter transformer (since for ferrites, we can have B^2 dependency in the core-loss equation). However, we are also using a (Flyback) core size that is twice that in a Forward converter. So it is better exposed to cooling. But at the same time, everything else is scaling too. For example, we first calculate core loss per unit volume and then multiply that with volume to get the total core loss. So if volume is doubled, for the same flux density swing, we will get double the core losses! And so on. The picture is really murky. We do need to depend a lot on industry (and our own) experience here. In the case of this author, it was 400 cmil/A_{COR}, just for achieving Class A transformer certification (and barely so). So it is probably best to target 350 A_{RMS}/cm^2 at worst. A lower density is even better (say 250 A_{RMS}/cm^2). But what do others say?

- AN-4140 from Fairchild allows 500 A_{RMS}/cm^2, suggesting up to 600 A_{RMS}/cm^2.
- Texas Instruments, www.ti.com/lit/an/slua604/slua604.pdf allows 600 A_{RMS}/cm^2.
- International Rectifier, www.irf.com/technical-info/appnotes/an-1024.pdf, suggests 200 to 500 cmil/A_{RMS}. This translates to 400 to 1000 A_{RMS}/cm^2.
- AN017 from Monolithic Power allows 500 A_{RMS}/cm^2.
- AN-9737 from Fairchild, www.fairchildsemi.com/an/AN/AN-9737.pdf, asks for 265 A_{RMS}/cm^2, very close to our conservative suggestion of 250 A_{RMS}/cm^2.
- On-Semi, www.onsemi.com/pub_link/Collateral/AN1320-D.PDF, allows 500 A_{RMS}/cm^2.
- Power Integrations recommends 200 to 500 cmil/A, *but in calculations often uses the COR value without necessarily pointing it out*, and typical values used are 500 A_{COR}/cm^2. That is 19,737/500 = 400 cmil/A_{COR}, same as what was suggested in the *A-Z* book. From page 333, that is 350 A_{RMS}/cm^2. See Fig. 10.4 too.

Keep in mind there is a big difference in making a small and *attractive* transformer for an evaluation board, and between a commercial product that meets safety approvals.

Comparison of Energy Storage Requirements in a Forward and Flyback

Irrespective of efficiency considerations, the most basic question is: By going from a Flyback to a Forward, do we end up requiring more magnetic volume or less?

We saw earlier, that when we went to the Flyback, its *transformer volume was twice that of the Forward converter*. But the Forward converter has an additional magnetic component, its energy storage element, that is, its secondary-side choke. Generally we pick an off-the-shelf inductor for that. But we can ask: If we use a gapped ferrite for the choke, how will its volume compare with the transformer of the Flyback? Keep in mind that in a Flyback, its transformer is also the energy storage element.

The answer to this is on page 225 of the *A-Z* book, where we show that for a Buck, the volume is $(1 - D) \times$ the volume of a Buck-boost, for the same energy, current ripple ratio, etc. so for a duty cycle of about 0.5, the volume of a Buck choke will be half that of a Buck-boost.

We learned that the transformer of a Forward is half the size of a Flyback, but then we need a secondary-side choke for it, equal to half the size of the Flyback transformer. The total gain or *loss is virtually zero*. Both the Forward and the Flyback need almost the *same total volume* of magnetic components. Yes in a Forward, the heat gets split into two components and their total exposed area is more than that of a single component of the same net volume. This is one of the reasons a Forward is preferred at higher powers. But the Flyback also suffers from zener clamp dissipation and high-RMS output capacitor current.