Wire Table

AWG WIRE SIZE	RESISTANCE (ohms/foot)	WIRE AREA* Circular mils	WIRE AREA* cm^2 ($\times 10^{-3}$)	CURRENT CAPACITY (amperes) (1)	CURRENT CAPACITY (amperes) (2)
8	.00063	18,000	91.2	22.0	44.0
9	.00079	14,350	72.7	17.5	35.0
10	.00100	11,500	58.2	13.8	27.6
11	.00126	9,160	46.4	11.0	22.0
12	.00159	7,310	37.0	8.71	17.4
13	.00200	5,850	29.6	6.91	13.8
14	.00252	4,680	23.7	5.48	10.9
15	.00318	3,760	19.1	4.35	8.70
16	.00402	3,000	15.2	3.44	6.88
17	.00505	2,420	12.2	2.74	5.48
18	.00639	1,940	9.83	2.17	4.34
19	.00805	1,560	7.91	1.72	3.44
20	.01013	1,250	6.34	1.37	2.74
21	.0128	1,000	5.07	1.083	2.17
22	.0162	810	4.11	.853	1.71
23	.0203	650	3.29	.681	1.36
24	.0257	525	2.66	.539	1.08
25	.0324	425	2.15	.427	.854
26	.0410	340	1.72	.337	.674
27	.0514	270	1.37	.269	.538
28	.0653	220	1.11	.212	.424
29	.0812	180	.912	.170	.340
30	.104	144	.730	.133	.266
31	.131	117	.593	.1056	.211
32	.162	96.0	.487	.0853	.171
33	.206	77.4	.392	.0672	.134
34	.261	60.8	.308	.0530	.106
35	.331	49.0	.248	.0418	.0836
36	.415	39.7	.201	.0333	.0666
37	.512	32.5	.165	.0270	.0540
38	.648	26.0	.132	.0213	.0426
39	.847	20.2	.102	.0163	.0326
40	1.07	16.0	.081	.0128	.0256
41	1.32	13.0	.066	.0105	.0210
42	1.66	10.2	.052	.00833	.0166
43	2.14	8.4	.043	.00645	.0129
44	2.59	7.3	.037	.00533	.0107
45	3.35	5.3	.027	.00412	.00824
46	4.21	4.4	.022	.00330	.00660
47	5.29	3.6	.018	.00259	.00518
48	6.75	2.9	.015	.00233	.00466
49	8.42	2.25	.011	.00200	.00400
50	10.58	1.96	.010	.00130	.00260

* Based on maximum diameter of heavy Formvar wire with insulation.
(1) Based on 750 cir mil/amp. (2) Based on 375 cir mil/amp. Current capacity will vary according to the geometry and wire size, and can range from 375 to 1000 circular mils per ampere.

Figure 5.15 Typical vendor's table for selecting a wire gauge. (From Ref. 2, p. 13.)

> **Practical Note** When the manufacturer doesn't give the length per turn for 100% fill factor, or (more commonly) doesn't tell you what fill factor the length per turn is for, a good approximation for all core sizes can be made as follows: length per turn = *OD* + (2*Ht*), where *OD* is the unwound core outer diameter, and *Ht* is the unwound core height.

The catalog (Figure 5.12) lists the length per turn as 0.072 ft. The resistance per length of #22 wire, again from Figure 5.15, is 0.0162Ω/ft. The resistance (at 20°C) is thus

$$\begin{aligned} R &= \text{length/turn} \times \text{number of turns} \times \text{resistance/length} \\ &= 0.072\text{ft} \times 29 \text{ turns} \times 0.0162\Omega/\text{ft} = 34\text{m}\Omega \end{aligned}$$

about half the 75mΩ we initially calculated as absolute maximum permissible.

Power Loss

So far we've calculated the DC flux density and the resistance. To find total power loss in the inductor (aside from temperature, which will be done iteratively, see below), we still need the AC flux density, which determines the core losses. Let's calculate this next.

Recall that our switching frequency is 250kHz, which is a period of 4μs. The duty cycle was 33% (because $V_{out}/V_{in} = 5\text{V}/15\text{V} = 33\%$), and so the peak-to-peak ripple current was 0.377A.

So the core has a peak-to-peak AC flux density of $H_{AC} = H/NI \times N \times I_{AC} = 0.467 \times 29 \text{ turns} \times 0.377\text{A} = 5.1$ Oe. Since the permeability is $125 \times 80\% = 100$ (because the permeability has been reduced by the 2ADC current), the AC core flux density is $B_{AC} = H_{AC} \times \mu = 5 \text{ Oe} \times 100 = 500\text{G}_{pp}$.

Following the theoretical discussion above, we cannot find out what the core losses for this situation are really because the current waveform is triangular, not sinusoidal. Still, since all we have is losses for sinusoids, we're going to go ahead and get an approximate idea of the core losses by approximating the triangular waveshape with a sinusoid of the same peak-to-peak amplitude.

> **Practical Note** This approximation is one of the main reasons it's necessary to go to the lab and really measure your magnetics. It simply isn't possible to get really good power loss calculations for the core (you'll be doing very well if you're within 10–20%). Note however, that you can do much better when the magnetic piece really is a DC inductor, because if the AC ripple is zero, then so is the core loss.

Referring now to Figure 5.16, another chart from the Magnetics catalog, we find that with a flux density of 500G_{pp} at 250kHz, there is a core loss of approximately 30W/lb. [*Note*: This is a pretty crazy unit, huh? Other manufacturers give it in W/m^3.] The core has a weight of 0.0046 lb, so the core losses are about 140mW.

To get the total losses for the design, let's add in the copper loss: $P = I^2R = (2\text{A})^2 \times 34\text{m}\Omega = 136\text{mW}$ (at 20°C). Notice that the copper losses are just about the same as the core losses. Following our rule for optimal design, this means we have done a good job. If the ripple had been much smaller, yielding a smaller core loss, this would have told us to decrease the copper loss at the expense of increasing the core loss, which we would accomplish by removing turns, using a higher perm core, and letting the inductance swing more—just the direction we initially started from. In practical terms, this would mean that our restriction of not letting the inductance swing down to less than 80% of its initial

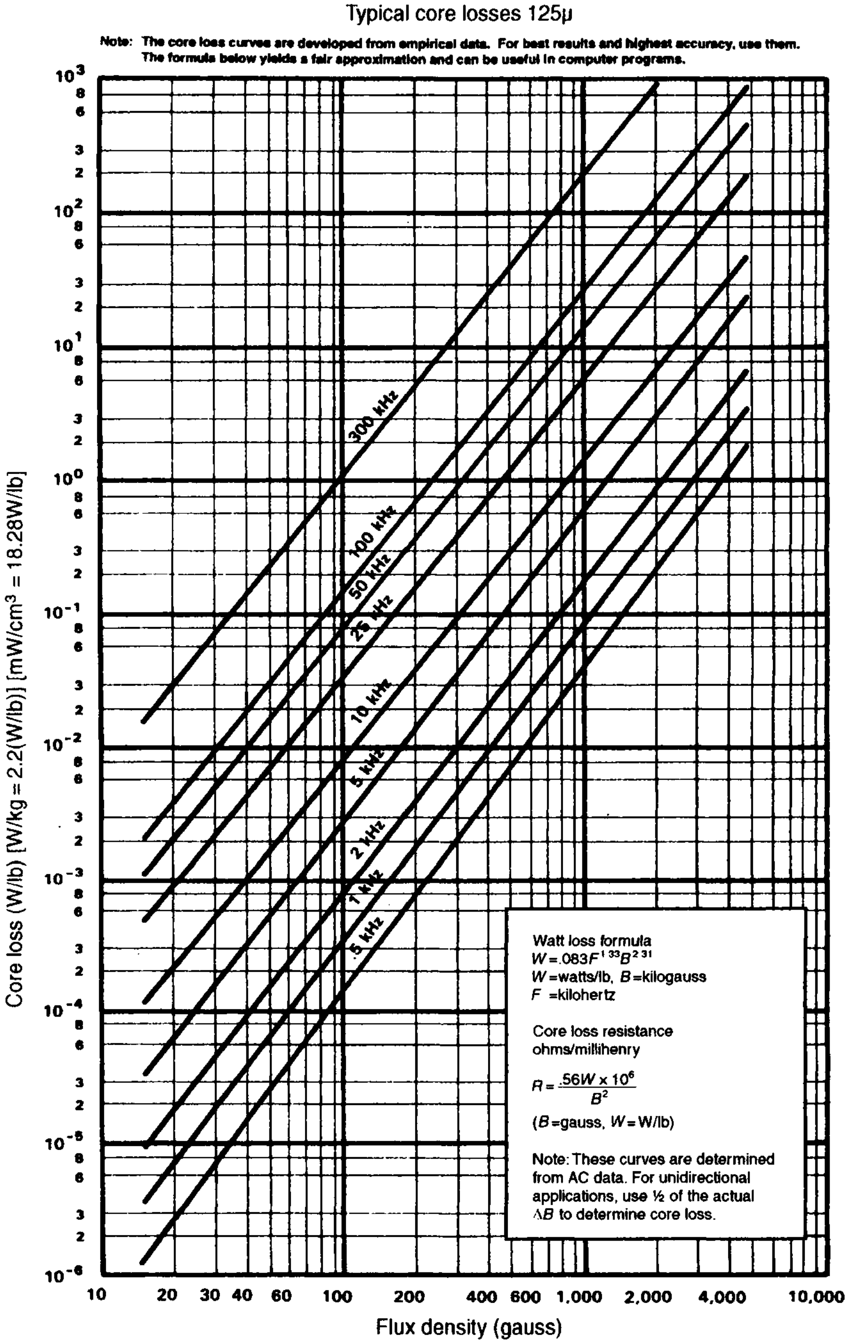


Figure 5.16 Calculating core loss. (From Ref. 2, p. 22.)

value would have been causing unnecessary losses. Of course, there may be a perfectly valid reason for limiting the swing, which in the case of a flyback is to prevent the pole set by the inductance from moving too much, as will become clear in questions of stabilizing the loop in the following chapter. At any rate, even though the losses are optimized already, at this point in the calculation it won't hurt to revisit this percent swing limitation to see if it can be usefully relaxed any.

The total power losses are $P = P_{\text{core}} + P_{\text{wire}} = 140\text{mW} + 136\text{mW} = 276\text{mW}$, and so we can estimate the temperature rise with the formula:

$$\Delta T = \left[\frac{(P(\text{mW})}{SA(\text{cm}^2)}\right]^{0.833} = \left(\frac{276}{2.5}\right)^{0.833} = 50°\text{C}$$

where SA is the (wound) surface area of the inductor, which is listed in the Magnetics catalog [2, p. 4]. If the power loss and thence temperature rise were dominated by the copper losses, and if the temperature rise had proven to be excessive, we would also have been pushed in the direction of fewer turns. Realistically, in this case it might be necessary to go to the next core size up. Conversely, if the temperature rise seemed acceptably low, the entire calculation could be repeated for the next core size down, to see if a smaller (and therefore cheaper) core would work.

Temperature Dependence

With the 50°C rise, we can now calculate the copper losses as a function of temperature. (Remember that the 136mW was based on the assumption that the wire was at 20°C.) The goal here is to produce a final power loss and temperature rise estimate that is self-consistent. That is, we want the temperature at which the power loss is calculated to be the same as the temperature at which that power loss implies the core is going to be operating. The equations governing the two equations are transcendental, so they can't be given a convenient form; but in reality, for all practical designs the temperature can be solved for iteratively, in just one or two cycles. Let's do that here, so you can see an explicit example.

The temperature rise calculated by our first estimate, ΔT, was 50°C. So the resistance should be multiplied by a factor of $(1.0039)^{50} = 1.21$, because copper has a positive temperature coefficient of 0.39%/°C; that is, the resistance at $20°\text{C} + 50°\text{C} = 70°\text{C}$ is $34\text{m}\Omega \times 1.21 = 41\text{m}\Omega$. The new power loss is 165mW in the copper, which is 305mW total, for a temperature rise of 55°C. This is close to the temperature at which we calculated the copper resistance, and so the whole calculation is now self-consistent. If the core operates only at 25°C, this 55°C temperature rise is perfectly acceptable.

Practical Note In practical applications, however, it is often desirable to limit the magnetics temperature rise to about 40°C.

For example, if the inductor is going to operate in an ambient temperature as high as 70°C, the inductor will be getting up over 125°C, and so you need a cooler design. Don't forget about maximum ambient temperature when calculating the wire resistance, either!

Conclusion

As you see, there can be quite a bit of work involved in the design of even the simplest magnetic structure, a DC inductor. People who do such designs frequently tend to use computer programs. All the major manufacturers offer such design programs for their cores, although the software is of widely varying quality and usefulness—caveat emptor!

PRACTICAL DESIGN OF A FLYBACK TRANSFORMER

Although we're not using a flyback transformer in our design of a buck converter (obviously), we'll give an exemplary design for one, because a flyback transformer is halfway between an inductor and a transformer, as indicated above, and deserves it own treatment for clarity. The presentation of the design work will be slightly less detailed than that for the DC inductor, but only on the aspects that are truly the same. Note that the design is for an isolated flyback; however, the design of a nonisolated flyback's inductor would be almost the same, except for the absence of a secondary. Let's suppose the following design requirements: a 48VDC input (for simplicity, we'll assume there is no line variation) and desired power output of 10W. Switching frequency is 250kHz. You've allotted 0.2W for losses (based on total losses you can allow to meet the converter's efficiency requirements), so the transformer has to be 98% efficient (0.2W/10W = 2%). This sort of efficiency is going to give you a moderate sized piece of magnetics; if the transformer has to be smaller, the efficiency will go down.

You can design the primary of an (isolated, discontinuous conduction mode) flyback transformer with just these four pieces of information: power output, switching frequency, losses, and input voltage. (They are also sufficient for designing the inductor of a nonisolated flyback.) Note that nothing has been said about inductance! Inductance is determined by the other parameters, as will become apparent below.

Let's say you're using the PWM chip UC3845, (a moderately priced 8-pin device), so the maximum duty cycle is 45%. The choice of maximum duty cycle is going to be related to the decision of whether this flyback is going to operate in continuous mode or discontinuous mode; we'll calculate it below. Our goal is going to be discontinuous mode for this example.

Let's set one more design goal: the transformer should be low profile, perhaps because of height constraints. It turns out that transformer design is not as straightforward as inductor design; there are always quite a few different magnetic cores that could be used to achieve the same electrical parameters. In this case, other criteria must be used to choose a core, based on size, or cost, or something else.

Equations Governing the Flyback

Let's do some basics first. As described in the theoretical portion at the beginning of this chapter, when the switch attached to the flyback transformer primary is on, the primary is acting like an inductor. Thus, we have a voltage applied across the inductance of the primary, and that results in a current that ramps up for as long as the switch is on:

$$I_{pk} = \frac{V}{L} t_{on} = \frac{V \times DC \times T}{L} = \frac{V \times DC}{f \times L}$$

where DC is the duty cycle, f is the switching frequency, and $T = 1/f$ is the switching period. This equation is valid because we are designing a discontinuous mode flyback. Remember that the current in the primary looks like the sketch in Figure 5.17.

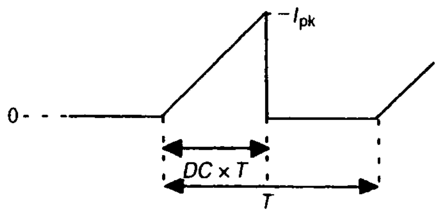


Figure 5.17 Current in a discontinuous mode flyback.

Now the energy stored in the primary inductance depends on the peak current:

$$E = \frac{LI_{pk}^2}{2} = \frac{V^2 DC^2}{2f^2 L}$$

and since this energy is delivered once every cycle,

$$P = Ef = \frac{V^2 DC^2}{2fL}$$

This equation is fundamental for the discontinuous mode flyback. It says that once the input voltage has been determined, to increase output power you have to either decrease the frequency or decrease the inductance; there are no other choices. Once the switching frequency has been chosen, all you can do to increase power is to decrease the inductance. Since there is a practical minimum to the inductance (set by, say, 10 times the stray inductances—let's say 5μH), there is a practical maximum amount of power you can get out of a discontinuous mode flyback converter, on the order of 50–100W.

> **Practical Note** At low input voltages above about 50W, you shouldn't be trying to design a flyback converter.

We've assumed that we're switching at 250kHz (perhaps set by switching transistor limitations). Calculating,

$$10\text{W} = \frac{(48\text{V})^2(0.45)^2}{(2 \times 250{,}000)L}$$

or $L = 93\mu\text{H}$. We can now find I_{pk}:

$$I_{pk} = \frac{48\text{V} \times 0.45}{250\text{kHz} \times 93\mu\text{H}} = 0.93\text{A}$$

Selecting a Core Material Type

Now we need to select a core material to achieve this inductance. Since the switching frequency is relatively high, we pick ferrite; another possibility would be MPP, and a full-blown design would properly consider that too, repeating all the steps herein. For simplicity of presentation, only the ferrite is considered, since it will probably turn out that the ferrite design is substantially smaller for the same efficiency than the MPP core design would be.

We already know that (in engineering units of centimeters, amps, and gauss)

$$B_{max} = \frac{0.4\pi I_{max} N \mu}{l_m} \tag{5.1a}$$

and

$$L = \frac{0.4\pi N^2 A_e 10^{-8} \mu}{l_m} \tag{5.1b}$$

with l_m the magnetic path length. Now for the small ferrite cores that we will be using, the magnetic path length is pretty tiny, with the result that B would be very large, probably saturating the core, and in any case certainly dissipating a lot of power. For this reason, *flyback transformers* (and any DC inductors that use ferrite) *always use an air gap*. The air gap greatly increases the effective magnetic path length because the permeability of air is so very much lower than that of ferrite. The effective path length for a core with an air gap is

$$l_e = l_m + \mu \times l_{gap} \tag{5.2}$$

In many practical cases it turns out that the second term of this equation is much larger than the first,

$$\mu \times l_{gap} \gg l_m \tag{5.3}$$

so that it is a reasonable approximation that

$$l_e \approx \mu \times l_{gap}$$

Note: This is only an approximation; it is not always true! You need to check that this approximation is true in every design, every time you use it.

Substituting in with this approximation, we have

$$B = \frac{0.4\pi I_{max} N}{l_{gap}} \quad \text{and} \quad L = \frac{0.4\pi N^2 A_e 10^{-8}}{l_{gap}} \tag{5.4}$$

Let's make it 100% clear about the usage of these equations: when there is an air gap in a ferrite (or other high perm) material, use equations 5.4, after verifying the validity of the approximation (equation 5.3); otherwise, use the fundamental equations shown in equations (5.1a) and (5.1b), remembering to use the effective path length (equation 5.2) when there is a very small air gap.

Core Selection

Unsurprisingly, it is usually necessary in actual practice to go through several different styles of cores to determine which is the best for a given application. The case we are designing for, however, had low transformer profile as a design criterion, which eliminates most styles from consideration. So we are going to go ahead and use the EFD style core (the name stands for "Economic Flat Design"); it would probably be reasonable to look at some other cores as well as this one when the design is finished, but we won't pursue that in the interest of space.

So let's pick the smallest EFD core, the EFD10, made, for example, by Philips [3], and see if we can squeeze the 10W out of it. If not, then we'll have to go up a core size. The information for this core is in the Philips soft ferrite cores catalog, reproduced as Figure 5.18.

Selecting Core Material

Now we can select a core material for this core. Referring to pages of the Philips catalog reproduced in Figure 5.19, we see that there are quite a few power materials from which to choose. In fact, if we look at other manufacturers' data books, there seems to be almost an endless variety, no two manufacturers making the same set of materials, not even materials with identical characteristics. How to choose?

Let's start by just talking about Philips's materials [1]. In the old days, everyone used a material referred to as 3C6A for everything in power. This material was pretty poorly characterized and had very high losses; it is now marketed as 3C80 and is used only in the most cost-sensitive applications. It was replaced by 3C8 material, which is now called 3C81. However, as converter switching frequencies continued to rise, Philips [1] came out with various new materials—remember that core losses grow faster than linearly with frequency. So nowadays, there is a whole set of power core materials, and we can pretty much choose one based on switching frequency alone.

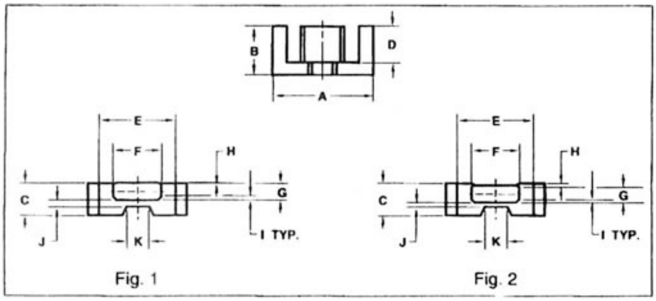
This also answers the problem presented by the availability of so many differing sets of core materials from each manufacturer. Closer examination will show that all have (at least roughly) similar materials for each frequency range, and indeed it is not uncommon in a magnetics specification to state that the core material used can be any one from a listed set, one from each manufacturer. Small differences in the materials are swallowed up by the tolerances of the various parameters in the construction of the magnetic core material.

Since we said that the switching frequency of this flyback was going to be 250kHz, we look across the soft ferrite materials selection table (Figure 5.19), and find that the recommended material is 3F3 (or, again, an equivalent from a different manufacturer). This material is very good, with losses half those of 3C85 at the same frequency; but things keep changing in this field, and you need to stay aware of the materials currently available. Perhaps there will be a better choice by the time you read this, but for our EFD10 core, we will choose 3F3 material.

Selecting the Gap

Having selected the core shape and material, we next select an air gap. The natural way to go about this might seem to be to target a peak flux density (based, e.g., on losses) and then

EFD Cores



Part Description* Example:

EFD 20 - 3F3

Material

Core Size (dimension A in mm)

Designates Economic Flat Design core Family

* Part Description is for a core half.

CORE DESCRIPTION	DIMENSIONS										
	A in/mm	B in/mm	C in/mm	D in/mm	E in/mm	F in/mm	G in/mm	H in/mm	I in/mm	J in/mm	K in/mm
EFD10 (Fig. 1)	.413 ±.012 10.5	.205 ±.004 5.2	.106 ±.004 2.7	.148 ±.006 3.75	.301 ±.01 7.65	.179 ±.006 4.55	.057 ±.002 1.45	.008 ±.004 0.2	.008 ±.004 0.2	.031 ±.004 .8	.118 REF 3.0
EFD12 (Fig. 2)	.492 ±.012 12.5	.244 ±.004 6.2	.138 ±.004 3.5	.179 ±.006 4.55	.354 ±.01 8.99	.213 ±.006 5.4	.079 ±.004 2.0	.008 ±.004 0.2	.008 ±.004 0.2	.031 ±.004 .8	.138 REF 3.5
EFD15 (Fig. 1)	.591 ±.016 15.0	.295 ±.006 7.5	.183 ±.006 4.65	.217 ±.010 5.5	.433 ±.014 11.0	.209 ±.006 5.3	.094 ±.004 2.4	.008 ±.004 0.2	.020 ±.004 0.5	.031 ±.004 .8	.138 ±.004 3.5
EFD20 (Fig. 2)	.787 ±.022 20.0	.394 ±.006 10.0	.262 ±.006 6.7	.303 ±.010 7.7	.606 ±.020 15.4	.350 ±.008 8.9	.142 ±.006 3.6	.007 ±.004 0.2	.031 ±.004 0.8	.031 ±.004 .8	.177 ±.008 4.5
EFD25 (Fig. 2)	.984 ±.026 25.0	.492 ±.006 12.5	.358 ±.008 9.1	.366 ±.010 9.3	.736 ±.024 18.7	.449 ±.008 11.4	.205 ±.006 5.2	.024 ±.004 0.6	.039 ±.004 1.0	.039 ±.004 1.0	.197 ±.008 5.0
EFD30 (Fig. 2)	1.181 ±.031 30.0	.591 ±.006 15.0	.358 ±.008 9.1	.441 ±.012 11.2	.882 ±.030 22.4	.575 ±.010 14.6	.193 ±.006 4.9	.030 ±.004 0.8	.039 ±.004 1.0	.039 ±.004 1.0	.236 ±.008 6.0

CORE DESCRIPTION	EFFECTIVE CORE PARAMETERS				A_L (nH) FOR UNGAPPED CORE SETS		
	le in/mm	Ae in²/mm²	Ve in³/mm³	C1 in⁻¹/mm⁻¹	3C85 ±25%	3F3 ±25%	3F4 ±25%
EFD10	.933 23.7	0.011 7.2	0.01 171	83.82 3.30	—	500	280
EFD12	1.12 28.5	0.018 11.4	0.02 325	63.5 2.50	—	700	380
EFD15	1.34 34.0	.023 15.0	.031 510	58.26 2.27	—	700	400
EFD20	1.85 47.0	.048 31.0	.089 1460	38.54 1.52	—	1150	580
EFD25	2.24 56.9	.090 58.1	.201 3300	24.89 1.00	2000	1800	
EFD30	2.68 68.0	.107 69.0	.287 4700	25.05 0.98	2100	1900	

Figure 5.18 A vendor's EFD core data sheet. For gapped core information, users are referred to pages 24 and 25 of Ref. 3; page 25 is reproduced below as Figure 5.21. (From Ref. 3, p. 18.)

Soft Ferrite Materials Selection Table

				FILTER INDUCERS			
PARAMETER[1]	SYMBOL	UNIT	TEST CONDITION	3B7	3B9	3D3	4C6
Suggested Freq. Range	f	MHz	—	< 0.3	< 0.3	< 2.5	< 20.0
Initial Permeability	μ_i	—	T = 25°C, B < 1mT	2300	1800	750	100
Saturation Flux Density @ Field Intensity	Bs @H	mT A/m	T = 25°C	410 250	350 250	370 1000	380 3000
Remanence	Br	mT	T = 25°C	150	125	160	200
Coercive Force	Hc	A/m	T = 25°C	15	20	80	250
Relative Loss Factor ($\times 10^{-6}$)	$\tan\delta / \mu_i$	—	f = 100 kHz, B < 0.1 mT f = 300 kHz, B < 0.1 mT f = 1 MHz, B < 0.1 mT f = 3 MHz, B < 0.1 mT	≤ 5 12	≤ 5 16	— ≤ 10 ≤ 30	— — 25 ≤60
Power Loss Density[2] @ T = 100°C (Sinewave Excitation)	Pv	mW/cm³	f = 25kHz, B = 200mT f = 100kHz, B = 100mT f = 500kHz, B = 50mT f = 1MHz, B = 30mT	—	—	—	—
Temperature Factor ($\times 10^{-6}$) in the Temp. Range	TF ΔT	K^{-1} °C	f = 10kHz, min max	- 0.6 + 0.6 +20 – +70	+ 0.9 + 1.9 -30 – +70	+ 0.5 + 2.5 +25 – +70	0 + 6.0 +5 – +55
Curie Temperature	Tc	°C	—	≥ 170	≥ 145	≥ 200	≥ 350
DC Resistivity	ρ	Ω.m	T = 25°C	≈ 1	≈ 1	≈ 2	≈ 10^5
Density	δ	g/cm³	—	≈ 4.8	≈ 4.8	≈ 4.7	≈ 4.5
Available Core Shapes	Planar E Cores		Page 5	—	—	—	—
	Pot Cores & PT Cores		Pages 6, 7, 8, 9, 10	●	●	●	●
	PQ Cores		Page 11	—	—	—	—
	RM Cores		Pages 12, 13	●	●	●	●
	E, EC, ETD Cores		Pages 14, 15, 16,17	—	—	—	—
	EFD Cores		Page 18 (Fig 5.18, this book)	—	—	—	—
	EP Cores		Page 19	—	—	—	—
	U&I Cores		Page 20, 21	—	—	—	—
	Toroidal (Ring) Cores		Pages 22,23	●	—	●	—

NOTES:
1. Values shown are based upon measurements on toroidal test cores with OD = 25mm, ID = 15mm, HT = 10mm. Products generally do not fully comply with the material specification. Deviations may occur due to shape, size, grinding operations etc.
2. Typical values. See detailed material specification for guaranteed power losses. Each core in our catalogs has a guaranteed power loss specification.
3. For detailed material information, request our Soft Ferrite Material Selection Guide, PC1062

Figure 5.19 Catalog pages showing characteristics of the vendor's soft ferrite materials. (From Ref. 3, pp. 2–3.)

determine a gap that gives this flux. (What is meant, of course, is that knowing both the flux density and the desired inductance is what determines the gap—of course the flux density alone is insufficient because of its dependence on the number of turns.) The problem with this approach is that it ends up with an odd gap size that will have to be specially ground for this transformer—read money. Another potential problem with making a selection this way is that the gap could end up being very small, in which case the tolerance on the gap could have a significant effect on the flux density achieved, and thence the losses; there might even be the potential for saturation of the core.

Soft Ferrite Materials Selection Table

WIDE BAND & PULSE TRANSFORMERS, EMI INDUCTORS				POWER TRANSFORMERS & INDUCTORS					MAGN. AMP.	
4C4/4C65	3E2A	3E25	3E5	3C80	3C81	3C85	3F3	3F4	3R1	SYMBOL
—	—	—	—	< 0.1	< 0.1	< 0.2	0.2–0.5	0.5–3.0	—	f
125	5000	6000	10000	2000	2700	2000	1800	900	800	μ_i
380 3000	410 250	380 250	360 250	500 3000	500 3000	500 3000	500 3000	450 3000	450 3000	Bs @H
200	100	100	80	150	100	150	150	150	340	Br
250	5	5	5	20	15	15	15	60	40	Hc
— — 40 ≤ 80	≤ 10 25 — —	≤ 25 ≤ 200 — —	≤ 75 (≤ 25 @ f = 30kHz) —	—	—	—	—	—	—	tan δ / μ_i
—	—	—	—	100 — — —	90 @ 50°C — — —	80 90 — —	— 50 180 —	— — 180 140	—	Pv
—	—	—	—	—	—	—	—	—	—	TF
≥ 350	≥ 170	≥125	≥125	≥ 200	≥ 210	≥ 200	≥ 200	≥ 220	≥ 200	Tc
≈ 10^5	≈ 0.5	≈ 0.5	≈ 0.5	≈ 1	≈ 1	≈ 2	≈ 2	≈ 10	≈ 1	ρ
≈ 4.5	≈ 4.8	≈ 4.9	≈ 4.9	≈ 4.8	≈ 4.8	≈ 4.8	≈ 4.8	≈ 4.7	≈ 4.7	δ
—	—	—	—	—	—	●	●	●	—	Planar E
—	●	●		—	●	●	●	●	—	P, PT
—	—	—	—	—	●	●	●		—	PQ
—	●	●	●	—	●	●	●	●	—	RM
—	●	●	—	●	●	●	●		—	E, EC, ETD
—	—		—	—	—	●	●	●	—	EFD
—	—	●	●	—	●	●	●	●	—	EP
—	●	●	—	●	●	●	●		—	U&I
●	●	●	●	—	●	●	●		●	T, RCC

Figure 5.19 *(Continued)*

> **Practical Note** It's not really practical to specify an air gap less than 10–20 mils (thousandths of an inch; i.e., 0.25–0.5mm) because the tolerance on the grinding is 1–2 mils (0.025–0.05mm). Below this value, your only safe bet is to buy a pregapped core for which the manufacturer guarantees an A_L rather than a gap size.

Even with a pregapped core, you have to worry about how much this gap will change when the two core halves are clamped together if the gap is too small: a glue will add to the gap length (especially if the glue thickness varies from unit to unit), and if you pot the core, it may expand. There are all sorts of problems; designing with a gap larger than 20 mils avoids many of them.

Practical Note When you buy a core set that has a given A_L, it frequently has one half gapped and the other half ungapped. Thus, for lab work, you can achieve A_L values equal to half those listed in the book by putting together two gapped halves. Of course, then you're stuck with a bunch of ungapped halves.

Practical Note When you build your own gapped core in the lab, a common "gotcha" occurs when you try to put spacers in each of the two outer legs of the core (e.g., with multiple layers of 2-mil Mylar tape) and make each spacer equal to the desired gap. You need to remember that the gap you calculate is the total air path length, which is the sum of the center post path and (either one of the) outer post paths. (There are two complete paths, one through either side of the structure.) Since putting gaps on the outside legs also creates a gap in the center post, the gap you put into each leg should be half of this (see Figure 5.20).

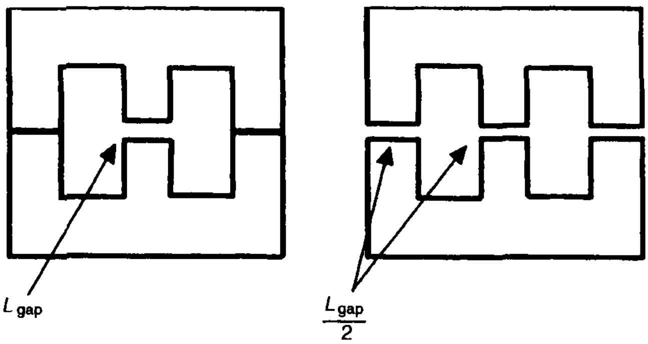


Figure 5.20 A center gap is equal to twice a side gap.

Practical Note If you want to have the equivalent of a 50 mil gap at the center post, you need to put spacers in the sides each of thickness 25 mils.

Returning to selecting a gap for our core, and looking at another page of the Philips catalog (Figure 5.21), we see that for the EFD10, there are five different A_L values available as standard products. Without thinking about it very deeply, we might suppose that 93μH seems like a lot of inductance on such a physically small core, so let's start by trying the core with the highest A_L. Since this implies it will have the fewest turns, it will also have the lowest winding resistance, which sounds promising. The highest A_L for this core is listed as 160nH. To get 93μH we need

$$N = \sqrt{\frac{93\mu\text{H}}{160\text{nH}}} = 24 \text{ turns}$$

Standard A_L Values For Gapped E Core Families

FOR 3C80, 3C81, 3C85, 3F3 MATERIALS

PART DESCRIPTION[1] New/(Old)	A_L[3] (nH) ±3% / μe				
E CORE SETS					
E13/6.4-"M"-A_	40	63	100	160	250*
(814E250PA_-"M")	44	69	109	175	273
E19/4.7-"M"-A_	40	63	100	160	250*
(813E187PA_)	56	89	141	225	352
E19/8.7-"M"-A_	63	100	160	250	315
(813E343PA_-"M")	48	77	123	192	242
E25/6.4-"M"-A_	63	100	160	250	315
(812E250PA_-"M")	64	102	163	255	321
E25/7.2-"M"-A_	100	160	250	315	400
(New)	89	142	222	280	356
E30/6.9-"M"-A_	100	160	250	315	400
(782E272PA_-"M")	91	145	227	286	363
E31/9.4-"M"-A_	160	250	315	400	630
(New)	95	148	187	237	373
E34/9.3-"M"-A_	160	250	315	400	630
(E375PA_-"M")	109	170	214	272	428
E41/12-"M"-A_	160	250	315*	400*	630**
(E21PA_-"M")	66	104	131	166	262
E42/15-"M"-A_	250	315	400*	630*	1000**
(783E608PA_-"M")	108	136	172	271	431
E42/20-"M"-A_	250	315	400	630*	1000**
(783E776PA_-"M")	84	105	134	211	334
E47/16-"M"-A_	250	315	400	630*	1000**
(E625PA_-"M")	76	95	121	191	302
E50/15-"M"-A_	250	315	400	630*	1000**
(New)	106	133	169	267	424
E55/21-"M"-A_	315	400	630	1000*	1600**
(E55PA_-"M")	88	112	176	280	448
E56/19-"M"-A_	315	400	630	1000*	1600**
(E75PA_-"M")	79	101	158	251	402
E65/27-"M"-A_	400	630	1000	1600*	2500**
(E65PA_-"M")	88	138	219	350	547
E80/20-"M"-A_	315	400	630	1000*	1600**
(New)	118	150	236	374	598
E83/38-"M"-A_	630	1000	1600*	2500**	3150**
(New)	121	193	308	482	607
PLANAR E CORE SETS (E CORE WITH PLATE)					
EI14/5.0-"M"-A_	25	40	63	100	160
(New)	23	37	58	92	148*
EI18/6.0-"M"-A_	63	100	160	250	315
(New)	26	41	65	102	129
EI22/8.2-"M"-A_	160	250	315	400	630
(New)	42	66	83	106	166
EI32/9.6-"M"-A_	160	250	315*	400*	630**
(New)	35	55	69	87	138
EI38/12-"M"-A_	250	315	400	630*	1000**
(New)	45	57	72	113	180
EI43/14-"M"-A_	250	315	400	630*	1000**
(New)	45	57	72	113	180
EI58/15-"M"-A_	315	400	630*	1000*	1600**
(New)	56	71	112	178	285

Part Description Example:

E 30/6.9 - 3F3 - A100

- A_L value (nH) or GAP size (μm)
  A-unsymmetrical gap
  E-symmetrical gap
  G-mechanical gap
- Material
- Dimensions in mm
  A/C[†] FOR E, I, U & T cores
  A/2xB for P, PT, PTS, PTR & PQ cores
  A for EC, EFD, EP and ETD cores
- Core shape

[†]For EI sets: "B" of E core + "C" of PLT core

PART DESCRIPTION[1] New/(Old)	A_L[3] (nH) ±3% / μe				
EC CORE SETS					
EC35-"M"-A_	63	100	160	250	315*
(EC35PA_-"M")	46	73	117	184	231
EC41-"M"-A_	63	100	160	250	315
(EC41PA_-"M")	37	59	94	147	186
EC52-"M"-A_	63	100	160	250	315
(EC52PA_-"M")	29	46	74	116	147
EC70-"M"-A_	160	250	315	400	630*
(EC70PA_-"M")	65	102	130	165	256
EFD CORE SETS					
EFD10-"M"-A_	25	40	63	100*	160*
(New)	66	105	165	263	420
EFD12-"M"-A_	40	63	100	160*	250**
(New)	80	125	199	318	497
EFD15-"M"-A_	63	100	160*	250*	315**
(New)	114	181	289	452	567
EFD20-"M"-A_	63	100	160	250	315*
(New)	76	121	194	302	381
EFD25-"M"-A_	100	160	250	315	400
(New)	80	127	199	251	318
EFD30-"M"-A_	100	160	250	315	400
(New)	78	125	195	246	312
ETD CORE SETS					
ETD29-"M"-A_	100	160	250	315	400
(ETD29PA_-"M")	75	120	188	237	300
ETD34-"M"-A_	100	160	250	315*	400
(ETD34PA_-"M")	65	103	161	203	258
ETD39-"M"-A_	160	250	315	400	630*
(ETD39PA_-"M")	94	147	185	235	370
ETD44-"M"-A_	160	250	315	400	630*
(ETD44PA_-"M")	75	117	148	187	295
ETD49-"M"-A_	250	315	400	630*	1000*
(ETD49PA_-"M")	106	134	170	267	425
ETD54-"M"-A_	250	315	400	630*	1000**
(New)	90	114	145	228	361
ETD59-"M"-A_	315	400	630	1000*	1600**
(New)	95	120	190	301	481

NOTES: * means A_L tolerance = 5% ** means A_L tolerance = 10%

1. Substitute selected material for "M" and fill in an A_L value after A in part description.
2. RM6R is also available in 4C6 with A_L=63, μe=41.
3. A_L measured @ f=10kHz, B≤1mT, T=25°C, Fill factor ≥ 80%, Clamping pressure ≡ 1N/mm² (≡ 150PSI).

Figure 5.21 Off-the-shelf pregapped EFD cores: vendor's table of A_L values. (From Ref. 3, p. 25.)

Note: The gap can be calculated by looking up $A_e = 0.072\text{cm}^2$, so that

$$160\text{nH} = \frac{0.4\pi\,(1\text{ turn})^2\, 0.072 \times 10^{-8}}{l_g}$$

which yields gap = 0.0057cm = 2.2 mils—tiny! Clearly, this is not the sort of gapping you should try to achieve on your own.

Knowing the gap, we can now find the flux density,

$$B = \frac{0.4\pi \times 0.93\text{A} \times 24\text{ turns}}{0.0057\text{cm}} = 4970\text{G}$$

which is greater than the saturation flux density of the core at 100°C of 3300G. (Although on the other hand it just squeaks by under the saturation flux density at 25°C of 5000G—conceivably you could be fooled in the lab!)

Continuing through the available options with the same calculations we find the set of values listed in Table 5.5. The last ($A_L = 25$nH) is the largest gap pregapped core Philips offers. Of this list, only the last two have flux densities less than the 100°C saturation flux density of 3F3 of 3000G, so we won't consider any further the cores with $A_L = 63$ and 100nH.

TABLE 5.5 Calculating Flux Densities of Pregapped Cores

A_L (nH)	N	l_g (cm)	B (G)
100	30	0.0090	3848
63	38	0.0144	3070
40	48	0.0226	2463
25	61	0.0362	1956

Core Loss

How about core loss for our two choices, $A_L = 25$ and 40nH? In a flyback, as shown at the beginning of the chapter, current is unidirectional, and therefore so is flux density: it increases from 0 to B_{max} and then back to zero, so that the peak to peak flux density is half of B_{max}. For the 3F3 material at 250kHz, losses at 2463G/2 = 1231G are approximately 330mW/cm^3; at 1956G/2 = 978G they are approximately 170mW/cm^3. (The Philips catalog also describes 3F3 characteristics: see Figure 5.22).

How Did He Read That Little Graph?

No, the author can't read tiny little graphics any better than you can—the trick is to write an equation of the form $\text{mW/cm}^3 = a \times B^x$, where a and x are constants, and determine their values by selecting two points that cross axis lines exactly so you can read their values well. There are then two equations in two unknowns, easily solved by hand or with a math program.

To be specific, for 3F3 material at 200kHz, we can pick 500G, where the losses are 20mW/cm^3, and 800G, with 80W/cm^3. The two equations are:

$$20 = a500^x$$

$$80 = a800^x$$

1 SERIES 3F3 FERRITE MATERIALS

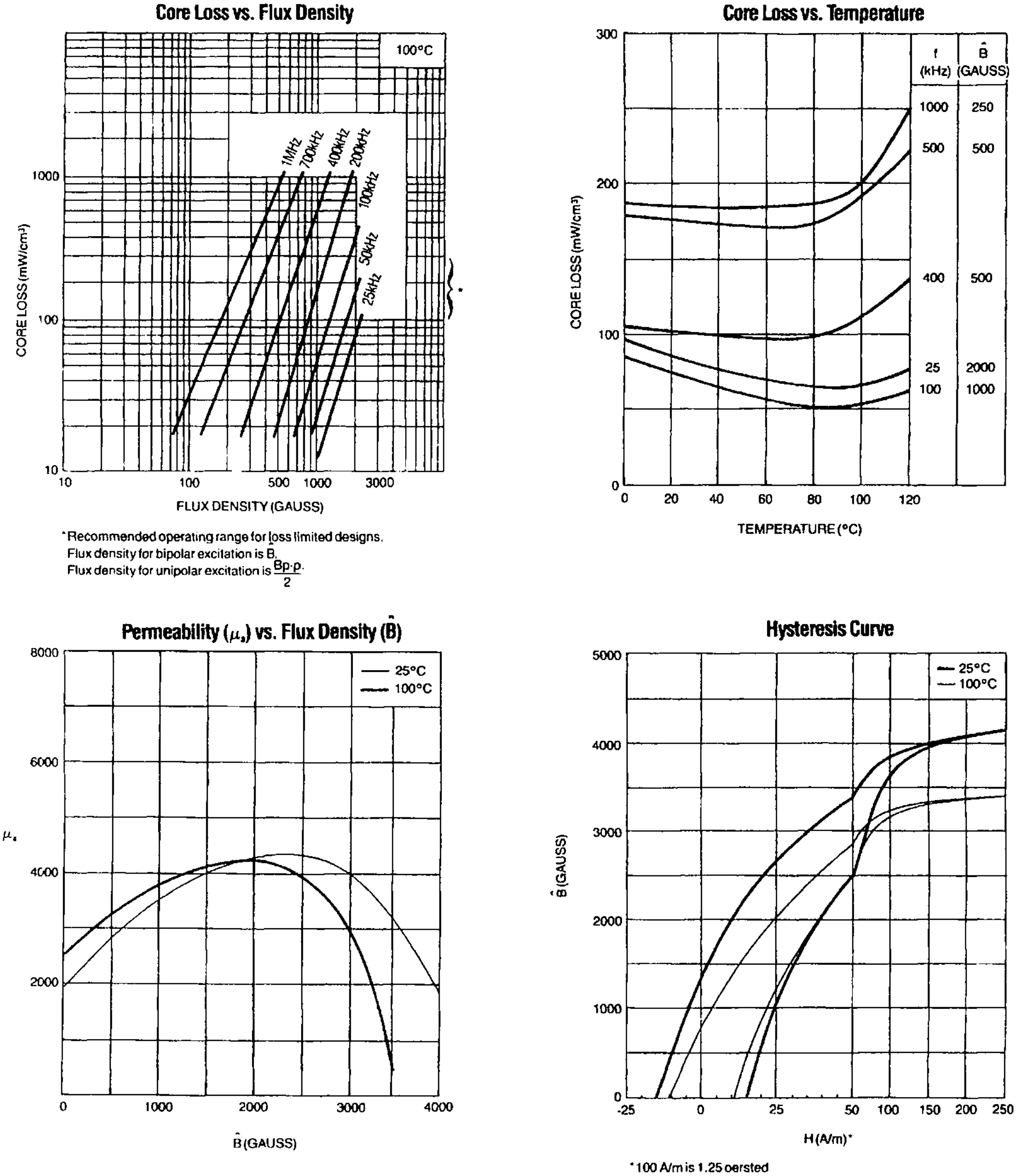


Figure 5.22 Vendor's presentation of 3F3 characteristics. (From Ref. 1, p. 37.)

Multiplying the first equation by 4 on both sides gives

$$80 = 4a500^x$$

which combines with the second equation to give

$$4 \times 500^x = 800^x$$

Taking logarithms of both sides, we have

$$\ln(4) + x\ln(500) = x\ln(800)$$

which at once solves as $x = 2.94$. Substituting back into the original equation, $a = 2.19 \times 10^{-7}$. Thus, at 200kHz,

$$\text{mW/cm}^3 = (2.19 \times 10^{-7})B^{2.94}$$

Rather than try to interpolate based on frequency, we'll get into the right ballpark at 250kHz by simply multiplying this by a factor of (250kHz/200kHz) = 1.25, which is the source of the preceding estimates.

Can I Get Lower Core Losses by Lowering the Switching Frequency?

To answer this question, recall from the theoretical part of this chapter that losses depend nonlinearly on both frequency and flux density. A typical relationship might be

$$\text{losses/lb} = f^{1.2}B^{2.3}$$

So, for instance, let's see what happens if the switching frequency is cut in half:

$$f \to \frac{f}{2}$$
$$L \to 2L$$
$$N \to \sqrt{2}N$$
$$B \to \sqrt{2}B$$

because, respectively, we double the inductance to maintain the power level; which means root 2 times the number of turns to double the inductance; which increases B by root 2, because B is proportional to the number of turns.

Total losses, which are losses per pound times weight, therefore go as

$$\left(\frac{f}{2}\right)^{1.2}(\sqrt{2}B)^{2.3}(2L) \approx 1.92fBL$$

because core weight is directly dependent on the energy stored, which is linear in inductance. Thus core losses have almost doubled with a cut in half the switching frequency. On the other hand, a lower switching frequency does decrease switching transistor losses proportionately to frequency:

$$P_Q \approx K + Af$$

where K is set by the on-state losses and A by the switching speed. Therefore,

$$P_Q\left(\frac{f}{2}\right) \approx 0.5P_Q(f)$$

if the switching losses dominate over the on-state losses (as will be true at fairly high switching frequencies). The moral of the story is that in a typical situation, changing the switching frequency doesn't have huge effects on efficiency, though there may be an overall broad optimum to be found. The real benefits will be seen in the size of the magnetics, which decreases with increasing frequency.

Returning to the core losses calculation, the total volume of the core is $171\text{mm}^3 = 0.171\text{cm}^3$. Thus for losses for the first of the two cores we have $330\text{mW/cm}^3 \times 0.171\text{cm}^3 = 56\text{mW}$, and for the second, $170\text{mW/cm}^3 \times 0.171\text{cm}^3 = 29\text{mW}$. Total losses, you recall, were supposed to be only 0.2W, so this seems to be working nicely so far; let's pick the lower A_L core for our design.

Had the core losses been unacceptably high, we would have two choices: either try to increase the gap still further, by mating two ground pieces or with a custom gap, or go on up to the next size core. As the gap gets larger, though, we start to have significant fringing (the magnetic field couples through the air out of the magnetic structure), which is to say there is increased leakage inductance. The increased leakage inductance will start to contribute to losses in the other elements of the circuit, negating the benefits we thought we were getting with the more efficient transformer design. On the other hand, a larger core takes up more board area, and costs more. As always in engineering, there are trade-offs to be evaluated.

Winding Losses

We can now calculate the copper losses for this design. This style core doesn't list the winding area, so let's figure it up directly from the specified core dimensions, given in Figure 5.23.

When calculating the winding area, remember that the wire goes in on one side and then back out the other side to complete the loop on the other side, so the winding area for half the core, as shown in Figure 5.23 (the whole unit consists of two of these pieces mated together) is the shaded area. Total winding area (WA) for this core is then double this,

$$WA = \frac{0.301\text{ in.} - 0.179\text{ in.}}{2} \times (0.148\text{ in.} \times 2) = 0.0181\text{ in.}^2$$

For a core of this shape, we may be able to achieve a fill factor as high as 80%. (If you need primary-to-secondary isolation, you had better count on substantially less fill factor: first allot the necessary area for the tape, and then use 80% for the wire in the remaining area.)

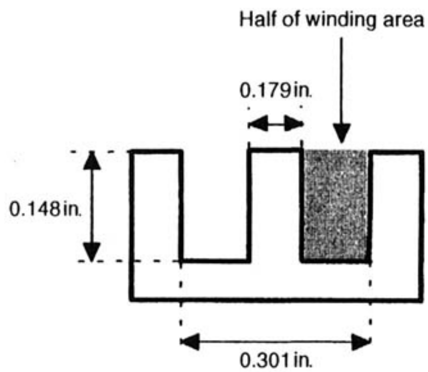


Figure 5.23 Calculating the winding area of an E core (not drawn to scale).

We can thus calculate the area per turn, remembering to use only half the winding area for the primary (so that we have half for the secondary):

$$\text{area/turn} = \frac{0.0181 \text{ in.}^2 \times 0.8}{(2)61 \text{ turns}} = 0.00012 \text{ in.}^2$$
$$= 28 \text{ AWG}$$

To get a conservative bound on the length per turn of the wire, let's assume that it goes from edge to edge of the winding area and is square in the third dimension (i.e., it would be bounded by a cube if removed from the core):

$$\text{length per turn} \approx 0.301 \text{ in.} \times 4 = 1.2 \text{ in. (conservative)}$$

So the resistance at 20°C will be not more than

$$R_{DC} = 1.2 \text{ in.} \times 61 \text{ turns} \times \frac{1000 \text{ ft}}{12{,}000 \text{ in.}} \times \frac{65.3\Omega}{1000 \text{ ft}} = 400\text{m}\Omega$$

The wire resistance of course goes up at higher temperatures. Supposing that the final magnetics temperature is 60°C (which can be figured out iteratively as was done in the DC inductor example above), the wire resistance will be

$$R = R_{20^\circ C} \times 1.0039^{(60^\circ C - 20^\circ C)} = 400\text{m}\Omega \times 1.0039^{40} = 467\text{m}\Omega$$

Usually, it is close enough to estimate the wire temperature from the ambient temperature and the power allotted for dissipation in the magnetic (using the surface area approximation demonstrated above). Otherwise, it can be done iteratively.

Do I Need to Worry About Skin Effect?

The skin effect causes current to flow in a sheath on the outside of a conductor. How deep the sheath is (the *skin depth*) depends on the frequency; at a low enough frequency, the skin depth is greater than the radius of the wire, in which case the entire cross-sectional area of the wire is being used. Thus at frequencies typical of switching power supplies, the skin effect can be important: doubling the cross-sectional area of a wire will not necessarily halve the resistance because the current stays on the outside of the wire.

On the other hand, going to multiple thin wires (litz) is not always a good idea either. Since each strand of the litz is individually insulated (if the strand weren't insulated, it wouldn't be an individual strand, it would be a funny-shaped solid wire), a lot of the winding area is potentially eaten up by the insulation. The number of strands that minimizes the resistance has to be decided on a case-by-case basis.

To decide whether to go with our design of 28 gauge wire, or use some kind of multiple-strand arrangement to decrease losses, we consider that the skin depth can be approximated by

$$\delta \approx \frac{6.61}{\sqrt{f}} \text{cm}$$

For our switching frequency of 250kHz, the skin depth is $\delta = 6.61/\sqrt{250{,}000} = 0.13\text{cm} = 0.0052$ in. Now for the 28 gauge wire we selected, the bare wire radius is 0.0063 in. (obviously the insulation thickness is irrelevant, since the material is nonconducting).

So the current-carrying cross-sectional area of the wire is the unshaded annulus in Figure 5.24, which has an area of

$$A = \pi[(0.0063\text{ in.})^2 - (0.0011\text{ in.})^2] = 0.000121\text{ in.}^2$$

Sometimes designers are advised to use wire smaller than the skin depth. Now what would happen if we did this and instead of a single #28 wire we used two strands of #31? (The wire scale is logarithmic, so increasing the wire gauge by 3 approximately halves the area.) Bare 31 gauge wire has a radius of 0.0044 in., which is less than the skin depth. Thus all the wire carries current, and the current-carrying area is $A = 2$ strands $\times$ $\pi(0.0046\text{ in.})^2 = 0.000133\text{ in.}^2$, about 10% larger than the effective cross section of the single strand of 28 gauge wire. But now let's include the wire insulation: the area of #28 wire with heavy insulation is 210 circular mils, and the area of two strands of #31, each with heavy insulation, is 2×110 c.m. $= 220$ c.m., about 5% larger than the single strand of #28. Thus, even ignoring questions of packing (two round wires don't fit as well as one round wire), you really aren't buying much of anything by going to multiple strands of smaller wire. *Don't assume that going to litz is buying you something; you have to check in each case.* In this case, we decide to stick with the single #28 wire.

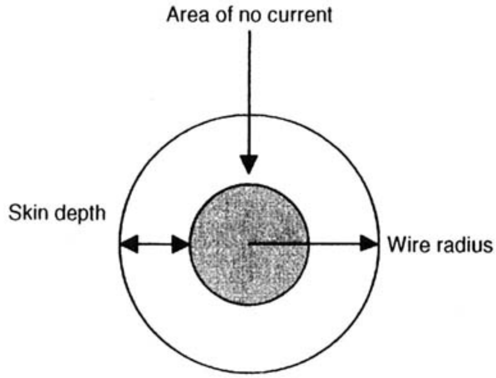


Figure 5.24 AC current only penetrates wire to the skin depth.

Copper Loss and Total Transformer Loss

Continuing with the evaluation of this design, remember that losses in the wire depend on the RMS current (Don't be confused on this one!) For the sawtooth current waveform shown earlier (Figure 5.17), the RMS current is

$$I_{\text{RMS}} = I_{\text{pk}}\sqrt{\frac{DC}{3}} = 0.93\sqrt{\frac{0.45}{3}} = 0.36\text{A}$$

Thus the power in the primary is $P_{\text{pri}} = (0.36\text{A})^2 \times 467\text{m}\Omega = 60\text{mW}$. Finally, since half the available winding area has been allocated to the primary, we may reasonably expect that the losses of the secondary will be equal to those of the primary, and we have the total power dissipated in the magnetic as $P_{\text{TOT}} = P_{\text{core}} + P_{\text{pri}} + P_{\text{sec}} = 29\text{mW} + 60\text{mW} + 60\text{mW} = 0.15\text{W}$. This is a transformer efficiency of 0.15W out of 10W, or 98.5%, meeting our original goal of transformer loss less than 0.2W.

Note that the copper losses are quadruple the core losses (0.12W vs. 0.03W). So, we really should be using fewer turns and a smaller gap; probably the $A_l = 40\text{nH}$ would be optimal. Since the design is already meeting spec, we won't pursue this any further.

FLUX DENSITY: TWO FORMULAS?

Up till now, we have been dealing with cores that store energy in themselves (and in their air gap), that is, inductors. (Recall that a flyback is an inductor during part of the switching period.) Now we are going to deal with transformers, magnetics that don't store energy. A brief digression is called for. Usually, people use different formulas for calculating the flux density in a transformer than in an inductor. This leaves you wondering where the formulas came from in the first place, and how does anyone know which to use when? This section will show that the two formulas are in fact identical, and the one selected is purely a matter of convenience, depending on which variables are known.

In engineering units, we already know:

$$L = \frac{0.4\pi \times 10^{-8} \times N^2 A_e \mu}{l_m} \tag{5.5}$$

$$B = \frac{0.4\pi \mu I N}{l_m} \tag{5.6}$$

and

$$V = \frac{LI}{t} \tag{5.7}$$

Let's rearrange (5.5) to solve for μ:

$$\mu = \frac{l_m L}{0.4\pi \times 10^{-8} \times N^2 A_e}$$

Substituting into (5.6) gives

$$B = \frac{0.4\pi I N}{l_m} \frac{l_m L}{0.4\pi \times 10^{-8} \times N^2 A_e} = \frac{10^8 IL}{NA_e}$$

but (5.7) is the same as $L = Vt/I$, so

$$B = \frac{10^8 I}{NA_e} \frac{Vt}{I} = \frac{10^8 Vt}{NA_e} \tag{5.8}$$

Thus, equations (5.6) and (5.8) are equivalent. Normally you use (5.6) for energy storage (inductors) because you know the current, and you use (5.8) for transformers because you are driving them with a voltage for a certain time; but these two formulas are equivalent, and give the same result for flux density.

PRACTICAL DESIGN OF A FORWARD TRANSFORMER

As an example of the design of a power transformer, we're going to design a forward, although again, we're obviously not using this in our buck design. Let's consider the following design requirements: we want a forward converter that has 48VDC in (for simplicity, we won't consider a range of input line voltages), 5VDC out at 100W, and a switching frequency of 250kHz. The basic configuration is shown in Figure 5.25.

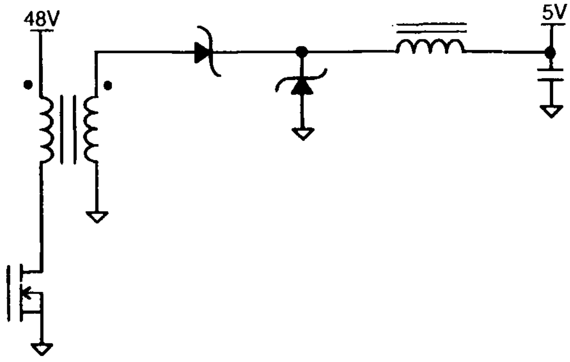


Figure 5.25 A forward converter.

Now the output current is $100\mathrm{W}/5\mathrm{V} = 20\mathrm{A}$. Since the current is high, we'll be using a small number of turns on the secondary, to keep the winding resistance low. In turn, this implies that the turns ratio (number of primary turns divided by number of secondary turns) for the smallest possible number of secondary turns, one, is going to be an integer. So to get started, let's see what happens if we start looking at integer turns ratios.

Turns Ratio = 1 : 1 This case has the same number of turns on the primary and the secondary. When the switching transistor turns on, the full 48V is applied across the primary, which in this case implies that 48V also appears across the secondary (ignoring leakage inductance), in turn applying it across the freewheeling diode. *Practically*, however, the highest voltage schottky diode that can be obtained that has reasonably low forward voltage is 45V. To use 48V will require at least a 60V part, and maybe higher if there is ringing, or if the input line has variation to it. This higher voltage diode will then have a higher forward voltage, which in turn will decrease the efficiency of the converter.

This question of rectifier diodes' forward voltage is always a problem for low voltage outputs. The reason is easy to see: the current through the inductor is always coming either through the rectifier diode or through the freewheeling diode; in either case, then, there is a loss of $V_f I$ through these diodes, and that is out of a total power of $V_{\mathrm{out}} I$, yielding an efficiency loss of V_f / V_{out} just from the diodes. The only way around this is to use synchronous rectifiers, but driving these is substantially more complex. (As V_{out} drops to 3.3V and lower, synchronous rectification becomes a necessity for just this reason.)

At any rate, for a reasonably high efficiency converter without synchronous rectifiers, a 1 : 1 turns ratio is not a good choice.

Turns Ratio = 2 : 1 Now the primary has twice the turns of the secondary, so that the 48V applied across the primary yields 24V across the secondary and the diodes, so a schottky can be used. The duty cycle of a forward converter is approximately $DC \approx V_{\mathrm{out}}/V_{\mathrm{sec}} = 5\mathrm{V}/24\mathrm{V} = 21\%$ (ignoring the V_{f} of the schottky.) The peak current on the primary, and thus through the switching transistor, may be calculated by recalling from the first part of this chapter that when the voltage steps up (secondary reflected to primary), the current steps down. So when there is 20A through the secondary forward diode, there will be $I_{\mathrm{pri}} = 20\mathrm{A}/2 = 10\mathrm{A}$ in the transistor. *Practically*, this may be too high for a MOSFET. (We won't be using a bipolar at 250kHz!) Since the MOSFET on-state losses go as the square of the current, the part will have 100 $A^2 \times R_{DS,on} \times 21\%$ losses, and a suitable FET may be too expensive to keep this loss to a reasonable level.

Turns Ratio = 3 : 1 Now the secondary diodes see only 48V/3 = 16V, and the duty cycle is about 5V/16V = 31%. The primary current is 20A/3 = 7A, so the on-state transistor losses are about three-quarters what they were at 2 : 1, only 49 $A^2 \times R \times 31\%$. All parameters seem to be under control for this turns ratio.

Turns Ratio = 4 : 1 The secondary diodes see only 48V/4 = 12V, and the duty cycle is up to 5V/12V = 42%. If you take into account the forward voltage of the diodes, or if the line can go lower than 48V, this will exceed 45%, which is the limit in duty cycle for PWM ICs such as the Unitrode UC3845. Thus *practically*, we have a limit from our choice of chips.

The conclusion from these calculations is that something like a 3 : 1 turns ratio best meets the various practical limits on components. Let's thus choose a 3 : 1 turns ratio.

Rather than going through the whole process of choosing a core, working through the gory details, seeing if some other core is better and so on, let's choose a suitable core to start with, assuming that all this other work has been done. Now we can concentrate on aspects of the problem that are novel in the design of the forward transformer.

Having said this, we choose an RM10 core with no center hole, which has an $A_e = 0.968\ \text{cm}^2$, and when using 3F3 material, has an $A_L = 4050\text{nH}$. With a three-turn primary, we have a primary inductance of $L_{pri} = (3\ \text{turns})^2 \times 4050\text{nH} = 36\mu\text{H}$, which results in a magnetizing current of

$$I_{mag} = \frac{48\text{V} \times 31\% \times 4\mu\text{s}}{36\mu\text{H}} = 1.6\text{A}$$

The RMS of this current is added RMS onto the primary current of 20/3A reflected from the secondary. We have

$$I_{RMS} = \sqrt{I_{DC}^2 + \frac{I_{pk}^2 DC}{3}} = \sqrt{\left(\frac{20}{3}\right)^2 + \frac{(1.6)^2 0.31}{3}} = 6.686\text{A}$$

resulting in an increase in loss, which is proportional to I_{RMS}^2, of $(6.686/6.66)^2 = 1.006$ or 0.6%, which although quite acceptable, will still be reduced a little bit for the sake of the discussion. To reduce the magnetizing current, we will increase the primary inductance, and so we increase the primary number of turns while maintaining the same turns ratio.

Selecting next a turns ratio of 6 : 2, the number of turns is doubled, so the primary has an inductance four times larger, 144μH, resulting in a peak magnetizing current four times smaller, 0.4A. This now gives a truly negligible increase in I_{RMS}^2.

Now we can also calculate the core flux density (remember that the 48V is applied for a time equal to the period times the duty cycle),

$$B = \frac{(48\text{V} \times 31\% \times 4\mu\text{s}) \times 10^8}{6\ \text{turns} \times 0.968\text{cm}^2} = 1025\,\text{G}$$

which seems to be a practical level to have limited losses with 3F3; note that the three turns tried originally would have resulted in a flux density of 2050G, which would have had very high core losses, one real reason for increasing the number of turns on the primary.

Now, just as in other designs, this design should go on to calculate core and copper losses, compare them with the next step of a nine-turn primary, and see which is most efficient. The various other steps proceed as before.

PRACTICAL DESIGN OF A CURRENT TRANSFORMER

As a final piece of magnetics design, we will design a current transformer, which could be used to reduce the losses in sensing the primary current in a converter.

What's the difference between a current transformer and a voltage transformer? This question seems to cause even experienced magnetics designers to scratch their heads. The fundamental difference can be expressed by saying that the voltage transformer is trying to reflect a voltage across from its primary to its secondary, whereas a current transformer is trying to reflect a current; the voltage that the current transformer sees depends on its load.

Working this through for a practical design case should make this clear.

Let's suppose for specifications that we want to sense the primary current on a converter, and to develop 1V for a current of 10A. Of course, we could just use a $1\text{V}/10\text{A} = 100\text{m}\Omega$ resistor, but this results in a loss of $1\text{V} \times 10\text{A} = 10\text{W}$, which is unacceptably high for almost all designs. So instead, let's use a current transformer arranged as in Figure 5.26.

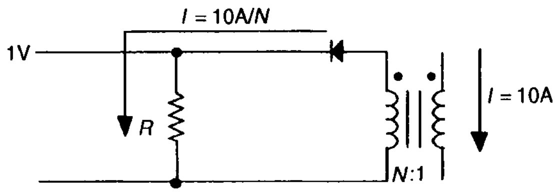


Figure 5.26 Using a current sense transformer to reduce losses.

Of course, we will use only one turn on the primary, to minimize the resistance, and many turns on the secondary, to drop the current down to a low level. If N is the number of turns on the secondary, by Ohm's law $(10/N)R = 1\text{V}$, and the power dissipated in the resistor is $P = (1\text{V})^2/R$. Let's suppose that we limit the power dissipation to 50mW (e.g., so we can use a derated 100mW resistor). This sets R to be no smaller than 20Ω, and using this value, Ohm's law shows that $N = 200$.

Now let's look at the core. If we suppose that the diode is a plain rectifier, we might expect a forward voltage of about 1V at a current of $10\text{A}/200 = 50\text{mA}$. So the total voltage the transformer sees is the 1V output, plus the 1V diode drop, or 2V. Then the flux density in the core, if we are operating at 250kHz, will be not greater than

$$B = \frac{(2\text{V} \times 4\mu\text{s})10^8}{200 \text{ turns} \times \text{A}_\text{e}} = \frac{4}{A_\text{e}}$$

since the time the current is passing through the primary can't be greater than the period (otherwise the core could never reset). Thus A_e can be quite small without making B very large, and thus the size of the core is not determined in this case by the need to limit losses or prevent saturation, but more likely by the separation between primary and secondary required for isolation voltage. If isolation isn't required, the core size is probably limited by the 200 turns: you may be able to carry a 50mA peak current in a #40 wire, but this gauge is so thin that vendors will refuse to wind it.

Practical Note Don't use wire gauges smaller than #36 unless you absolutely have to.

So now how do we know that our device isn't a voltage transformer instead of a current transformer? Consider that we have 2V on the secondary, and therefore $2V/200 = 10mV$ on the primary. If the source driving the current transformer is, for example, 48V, then the 10mV across the primary is insignificant—you can get the 50mA from the secondary without affecting the drive to the current transformer's primary. Suppose on the other hand (unrealistically) that the driving source on the primary were only 5mV. Then it wouldn't be possible to generate 10mV across the primary, and you thus wouldn't be able to get the 50mA out of the secondary because the primary impedance (i.e., the reflected secondary impedance) is too large and is in fact determining the current. Even if the entire 5mV were dropped across the primary, only $200 \times 5mV = 1V$ would be generated on the secondary: it can't produce enough voltage to drive the current through the resistor. Therefore it would act as a voltage transformer.

Viewed the other way, when the source is 48V, something other than the voltage on the primary is determining the current through the current transformer.

A current transformer is a voltage transformer that is impedance limited.

Finally, what about errors in the current transformer? The answer to this follows from the fundamental statement of what a current transformer is.

Practical Note The diode and the winding resistance of the transformer secondary don't matter to the measurement of the current, because (as long as it is impedance limited) the same current is going to flow through the resistor no matter what else is in series with it.

Practically, this is why it often doesn't matter whether you use a schottky as the rectifier: the lower forward voltage affects only the transformer, not the current measurement.

Measurement error does arise, however, if there is finite inductance, that is, magnetizing inductance. Suppose that we want to be sensing current with a maximum error of 1%. Since the secondary current is going to be 50mA, this means we have to have a magnetizing current (on the secondary) of less than $50mA \times 1\% = 500\mu A$. The magnetizing current diverts current away from the resistor, and thus we end up not measuring it, which is to say it is in error. We thus need to have a minimum inductance on the secondary of

$$L \geq \frac{2V \times 4\mu s}{50mA \times 1\%} = 16mH$$

With 200 turns, this means we need an A_L of $16\text{mH}/(200\ \text{turns})^2 = 400\text{nH}$, which is easily achievable with normal small ferrites.

TIPS FOR DESIGNING MANUFACTURABLE MAGNETICS

So far in this chapter we've presented theoretical aspects of magnetics, followed by practical guidelines for making a design that will repeatably work in the lab the way you want it to. But unlike most other electrical components, magnetics also have to be custom-produced in a factory, often one at a time. So after you've designed something that works in the lab to your satisfaction, the next step should be to talk to a manufacturer and try to make the unit work to *his* satisfaction as well. The best design is of no use if it can't be produced reliably, and so this section will give you some pointers based on many years' work with manufacturers.

Manufacturers of custom magnetics have a lot of experience, and you should listen carefully when they make suggestions on how to wind something or pot it, etc. Almost invariably, these suggestions have to do with minimizing the cost of production of the magnetic, which is of course highly desirable for your design. On the other hand, don't take a manufacturer's word uncritically, because someone who designs magnetics for a living is not necessarily well versed in circuits that use magnetics. In particular, be very cautious about suggestions for reordering the layers in a multiwinding transformer, because this strongly influences coupling. The usual answer to a request to change the order of the layers should be no, or at best, send a sample and try it out.

Wire Gauge

We've already talked about this, but re-iteration in the new context will be helpful.

Practical Note It's best to limit wire gauges to a maximum of #20, and a minimum of, say, #38. Above #20, some machines can't wind the wire, upping your cost, and above about #18, bobbins can be cracked by the stiffness of the wire. Try multiple strands of #20 if you need greater wire cross-sectional area. Below #38, manufacturers will of course wind wire, but it becomes very hard for you to build your own sample magnetic in the lab. The wire is like a piece of hair, subject to twists and snaps when you've just got that second-to-last turn on.... You may be better off using #38 even if you've calculated that that 1mA winding only needs #45, just because of the difficulty of handling.

As long as we're thinking about wire gauges, consider the possibility of saving money (if you're producing a lot of units) by controlling the number of different wire sizes used. If you have a primary winding using 23 gauge wire and two secondaries, one with #24 and one with #22, you might consider whether the design would still work if all three windings went to #23, or even #24. The cost saving might be quite noticeable, whereas an extra few milliwatts of loss might be more easily made up elsewhere.

Wire Gauge Ratio

The author has never heard anyone (except for the technicians who actually have to wind magnetics) mention this, but winding can become awkward if there are radically different sizes of wire on the same piece. The trouble is that if you wind some very thick wire in a flat spiral, and then try to wind some very thin wire on top of it, the thin wire tends to fall into the crevices between the turns of the thick wire, so that the thin wire doesn't form a flat layer. This can affect coupling, making it different from unit to unit. No firm guide is possible, but:

> **Practical Note** Try not to use wire sizes more than about 10 wire sizes apart on adjacent layers.

Toroid Winding Limits

Winding a toroid takes a lot more effort than winding magnetics of other types. Indeed, the author jokingly tells people that it can be proven that it is *topologically impossible* to wind a toroid! Anyway, the hand work is substantial, not to mention the very real and annoying prospect of losing count of how many turns you have put on. I always advise technicians to go to a place where they can't be interrupted, and make a mark on a paper for every 10 turns wound. Additionally, because winding is so labor intensive:

> **Practical Note** Cut your technician (or yourself) a break. Don't design a toroid with more than 200 total turns if you intend to hand-wind a sample.

Tape versus Wire Insulation

Tape is commonly used on a transformer to provide isolation voltage between a primary and a secondary, and sometimes for isolation between secondaries. There are two slightly different reasons here. Many safety agencies require a high-pot test between windings that are connected to an AC mains and windings that connect to where people can get at them. Depending on the circumstances, this test voltage can be anywhere from 500V to 3000V. This isolation is a perfectly natural use of tape, and at the upper end of the voltage range it may even be best served with a flanged bobbin—that is, one that has a piece of plastic dividing the winding area into two pieces, permitting the primary and secondary to be wound in separated compartments.

Isolation between secondaries differs in that it is not mandated by a regulatory agency, but is rather imposed by the designer to avoid arcing. For example, consider a flyback transformer that is generating a $+30$V and a -160V. In the same way that you keep the traces spaced apart to avoid arcing, the windings can't get too close together, either. There is about 200V difference between windings when the transistor is on, and possibly more when the transistor is off, depending on the details of the design: a forward has higher voltages inversely proportional to the duty cycle. Indeed, for a high voltage output such as the -160V, even individual layers of the wire may require insulation: you wind from left to right for one layer, than back from right to left for the next layer, and consider the maximum voltage from the underlying wire on the left side to the overlying wire on the same side.

While this second isolation requirement may also end up with tape, you should be aware that any layer of tape added to a design greatly increases cost because it is a hand operation. So for intralayer insulation, and also nonagency interlayer insulation, consider using heavier insulation on the wire, rather than tape. Standard insulation ("heavy") is a double layer, but both triple and quad are easily obtainable at not much cost increase, and they take less room than a layer of tape. The hard question is, How much insulation is enough for a given voltage?

Without getting into the details, the problem is that the voltage rating of magnet wire is given for 60Hz sinusoidal voltage, which is almost irrelevant for 100kHz square-wave operation, at which frequency little is known in any systematic way. Breakdown of the wire is also statistical, depending as it does on temperature and number of years of operation. Up to several hundred volts (peak), quad-build wire at switching–converter frequencies and waveforms appears to work fine. It should be good for almost all intrawinding insulation, and most nonagency interwinding insulation. This is the best I can tell you; the only way to be *sure*, is to run an accelerated life test in a real circuit.

Layering

The correct way of winding a multilayer winding has already been touched on: it should go left to right in one layer, then back right to left in another layer, and so on (this is not for a toroid, now). Although such a configuration is possible, consideration of the placement of the pins indicates that a winding should not end anywhere in the middle of a layer: if a winding started or terminated in the middle of a layer, it would have to cross over the rest of the layer to get to a side, where it could then exit the winding and make its way to a pin for connection. This crossover would be an uneven lump in the middle of the next winding on top, throwing it off. So part of the design-for-manufacturing process has to be selecting a wire gauge that enables you to get an exact integer number of layers; this consideration often dominates the desire to optimize resistance in the design of real magnetics.

Number of Windings

It is considerations of the kind just explored that compel manufacturable magnetics to have an absolute maximum of four to six windings. Not only is it difficult electrically to have more (because coupling becomes highly variable for the last couple of windings), but layering becomes difficult, and, bottom line, most bobbins have only 8–12 pins available! A custom bobbin is absolutely the last thing you want.

Potting

Potting is the process of filling up a volume surrounding a magnetic with a thermally conductive compound for the purpose of improving heat removal by providing a better thermal path, as well as by increasing the surface area of the magnetic mechanical structure. Potting is not to be confused with vacuum impregnation, which is used for insulation but doesn't do anything thermally.

Potting's big advantage is thermal and mechanical: it really helps get the heat out, and because it provides a flat surface, it can be very useful for mechanical mounting. (For example, a screw hole can be included in the potted shape.) There are also some potential problems with potting, which you should discuss with the vendor. The first is pretty obvious: potting compound is heavy, and your magnetic will weigh a lot more potted. Much less obvious, potting can change the magnetic characteristics of the magnetic. One problem much struggled with in the past is the shrinkage of potting material as it cures. This shrinking has been known to change the gap on gapped cores, causing the inductance to change! A similar problem is that ferrite cores, being rather brittle, can be snapped by the shrinkage. And a third problem along these lines is that MPP cores are strain sensitive, and their permeability can be affected by the shrinkage. There are various possible solutions for these problems, many of which revolve around proper selection of potting compound, but make sure your vendor is professionally dealing with these questions.

Specs

This last item is a peculiar one, really nothing that would occur to you as a reasonable designer—until you've experienced it a couple of times. It's quite challenging to write good magnetics specs. On your first couple of tries, you leave things out that you never imagined should be included (e.g., how far up the side of the bobbin should a layer of tape go?). Then on your next try, you put everything conceivable into the spec (which is now a 25-page book), and the vendor tells you that all this detail will cost you an arm and a leg.

You finally write something that satisfies you both, the vendor sends you a fourth sample, it works in your breadboard, the world is a good place. Now, your buyer finds a second source for the magnetic and sends them a copy of the spec to build to. They send you a sample, and it doesn't work at all! You take it apart to find out what's wrong, and you find that they've managed to misinterpret your masterpiece, building it to spec in a way that comes out completely different from your design.

This sounds like a horror story, but those with experience will recognize it as an every-supply occurrence. In fact, *the same vendor* can send you a sample built one way to a spec and then provide shipments to the same spec which are built another way! The only way the author has found that somewhat gets around these problems is to work with each vendor until something is produced that works, and then write into the spec that units must be built identically to the sample provided (number such-and-so). Constant vigilance is called for.

CONCLUDING COMMENTS

We thus see that there is nothing mysterious about (elementary) magnetics design. It just requires a lot of patience and attention to details—all the formulas really do work! But it is this need for patience and detailed work that is really the problem with magnetics design. If you design one or two magnetics once every other month, it perhaps is not too horrible to do it all by hand. But if you spend all day doing magnetics, and need to turn out two or three designs a day, it's not only tiresome, but impractical. The solution would be, of course, to let a computer do all the work. As suggested above, however, the available software doesn't seem to be adequate to the task. A worthwhile large-scale project for a group of engineers would be to design software that is technically accurate and has not only a modern user interface, but adequate documentation and a large enough database to be useful in designing at least the common types of magnetics covered in this chapter. The future awaits!

There are many advanced topics in the design of magnetics that were not addressed in this already long chapter. The author feels, however, that mastery of the practical techniques demonstrated here will suffice to generate most everyday magnetics designs. Following these steps should enable you to design a piece of magnetics that meets requirements the first time you build it, probably within 10–20%; this is the best that can be hoped for without very sophisticated and time-consuming analysis, which is done only for the most complex and critical designs. And to be truthful, the end result of many sophisticated analyses is still sometimes not as good as what can be accomplished by hand, based on the techniques in this chapter.

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