5

Practical Design of Magnetics

FUNDAMENTALS OF MAGNETICS

Introductory Comments

Designing a real piece of magnetics seems to take a lot of time. There are many decisions to be made: core material, core style, type of conductor, etc., and there won't always be a single best choice. When the design has been completed, performance has to be checked in the lab. And then you have to make sure the design can be produced at a reasonable cost!

Remember, though, that you can spend quite a bit of time designing even a window comparator: looking at resistor tolerances and temperature coefficients, leakage currents and offset voltages, and so on. So don't expect a magnetics design to jump into your lap either!

That said, though, this chapter gives a lot of practical information that will make your magnetics design, if not easy, at least straightforward. Although up to now knowledge of the basics has been assumed, in the field of magnetics, confusion among engineers is so common that it's best to begin with a review. The emphasis, however, remains strongly on the aspects of the theory that are essential for the practical design of magnetics. Indeed, after the introductory material, the rest of the chapter can be taken to be step-by-step instructions to producing good magnetics designs, including a final section about making your design manufacturable.

One last comment—This chapter cannot be a complete description of everything known about designing magnetics. Instead, it concentrates on aspects that are key for making a good, workable, solid design; to learn about proximity effect, field distributions and other advanced topics requires specialized study.

Ampere's Law

Let's start with the basics, then. All of magnetics is governed at a fundamental level by two laws, Ampere's and Faraday's.

Referring to Figure 5.1, imagine that we have N turns of wire wound onto a coil of circumference Length, and we're putting I amps of current through it. (The rectangle is just a former, something to hold the wire in place, plastic if you like; it will become a core later when we talk about permeability.) Ampere's law relates these parameters with the magnetic field, H, generated by the coil:

$$\text{Length} \times H = I \times N$$

By increasing the number of turns or the current, we can increase the field.

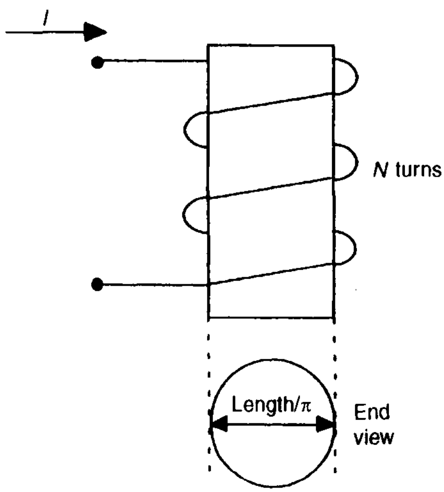


Figure 5.1 A coil of wire of N turns carries a current of I amps. The coil has a circumference of length equal to "Length".

Faraday's Law

The second law needed for magnetics is Faraday's. Referring to Figure 5.2, imagine that we have a loop of wire enclosing an area A with N turns on it. If the magnetic flux through that loop changes (e.g., by being coupled to another loop with a changing flux), then Faraday's law relates these parameters to the voltage developed at the terminals of the coil:

$$\frac{d\Phi}{dt} = \frac{d(BAN)}{dt} \rightarrow$$

$$V = NA\frac{dB}{dt}$$

where $V =$ rate of change of magnetic flux.

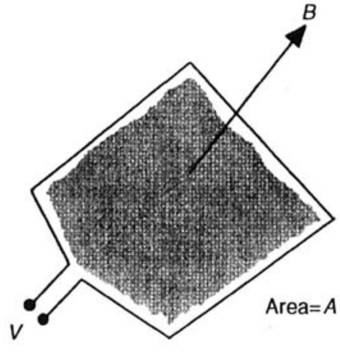


Figure 5.2 A loop of wire enclosing area A with voltage V and flux B.

About Inductance

From the fundamental laws of Ampere and Faraday, we can derive the equation governing the behavior of an inductor. First, there is a relationship between B and H. What are these two anyway? We called them both "the magnetic field" above without being very specific. In fact, it's only a historical accident that they have different symbols, since they're really the same thing: H is the magnetic field in free space, B is the magnetic field when it's inside a magnetic material. But they're exactly the same thing. The relationship between them is given by a property of the magnetic material, the permeability μ. You can think of μ as being the "gain" of the magnetic material, because it makes the field inside stronger than it would be if the material wasn't there:

$$B = \mu H$$

We can put this in Faraday's law (μ is assumed to be a constant in this elementary discussion), to get

$$V = NA\mu \frac{dH}{dt}$$

From Ampère's law we know that

$$H = \frac{IN}{\text{Length}}$$

so

$$V = \frac{N^2 A\mu}{\text{Length}} \frac{dI}{dt}$$

Inductance is now *defined* as

$$\text{Inductance} = L \equiv \frac{N^2 A\mu}{\text{Length}}$$

Notice the expected dependence of inductance on the square of the number of turns. With this definition, we have

$$V = L\frac{dI}{dt}$$

the familiar relationship expressing rate of change of current as determined by the voltage applied across an inductor.

Units Confusion

Engineers routinely use volts and amps and don't have to worry about dividing them: you get ohms. Unfortunately, with magnetics there are two systems, CGS (centimeter-gram-second) and MKS (meter-kilogram-second), and magnetics designs routinely mix units from these two systems. The reasons are again historical. So for magnetics, you have to carefully watch the units, and be sure to use conversion factors, such as 10 or 4π. (The 4π comes from units used in Maxwell's equations.) Table 5.1 details the units.

TABLE 5.1 MKS and CGS Units Often Used in Magnetics

	MKS	CGS
Field (H)	tesla (T)	oersted (Oe)
Field (B)	tesla (T)	gauss (G)
Inductance	henry (H)	square second per centimeter (s^2/cm)
Voltage	volt (V)	statvolt
Current	ampere (A)	electrostatic unit per second (esu/s)
Area	meter squared (m^2)	centimeter squared (cm^2)

Note how common it is to use gauss and henry in the same calculation! It's this that causes the confusion, since these units are from different unit systems. But who ever heard of a statvolt? And while the MKS system has the name tesla for the magnetic field, CGS has two different names, oersted and gauss. Table 5.2 shows how to convert between the two unit systems.

TABLE 5.2 Some MKS-to-CGS Conversions

MKS	(equals)	CGS
1 m^2		10,000 cm^2
1 volt		1/300 statvolts
1 amp		3×10^9 esu/s
1 henry		1.113×10^{-12} s^2/cm
1 tesla		10,000 gauss

Just so you'll be prepared, you should be aware that you sometimes see a few other units not listed above. "Amp-turn/cm" is another name for oersted [because Length (cm) times H (oersted) $= I$ (A) times N (turn)]. And as if all this weren't bad enough, sometimes *English* units are used! So note that 1 circular mil $= 5.07 \times 10^{-6}$ cm^2. Here, "mil" refers to one one-thousandth of an inch.

Let's note that μ is dimensionless, for example with B in gauss and H in oersted, because these are really the same thing: permeability is just a number. And finally, with area measured in square centimeters and length measured in centimeters, it ends up that inductance in henrys is

$$L = \frac{4\pi}{10} \frac{N^2 \mu A}{\text{Length}}$$

Weird Words: The Three R's

If you ever read the literature on magnetics, you will probably come across "the three R's," Reactance, Reluctance, and Remanence, which have done as much any anything in all of engineering to confuse people. Let's take a quick look at these R's, primarily just to demystify them.

Reactance This has similarity to resistance, it's a type of impedance. The reactance of an inductor is $\chi = 2\pi fL$ (the Greek letter is pronounced "kai"); similarly to Ohm's law, we have $V = I\chi$ for an inductor. Don't forget that reactance and resistance are 90° out of phase, so that the magnitude of the total impedance of a system that has both reactance and resistance is

$$|Z| = \sqrt{R^2 + \chi^2}$$

Remanence This is a property of a magnetic material, a type of magnetic hysteresis. Suppose you start putting a magnetic field through a magnetic material (e.g., by winding a coil around it and passing current through it—Ampere's law). Next, you reduce the applied field back to zero. *The magnetic material will still have a field in it.* The magnitude of this remaining field is called the remanence. Who cares? Well, the field in, for example, a transformer is cycled up and down through zero all the time; in this case, remanence is related to the amount of power dissipated in the material by the cycling. We'll discuss these core losses further below.

Reluctance This term is used in an analogy between magnetic and electric circuits—its purpose was supposedly to make magnetic circuits, which can be confusing, look like electric circuits, which everyone knows how to deal with. It seems to have grossly failed in this purpose! In this analogy, we define reluctance to be

$$\Re = \frac{1}{\mu A}$$

Using Ampere's law and the definition of flux that $\Phi = BA$, we can calculate the effect of a current through a multiturn loop:

$$I_{\text{tot}} = IN = \text{Length} \times H = \frac{\text{Length} \times B}{\mu} = \frac{\text{Length} \times \Phi}{\mu A} = \Re\Phi$$

Thus, if we say that Φ "looks like" voltage, then reluctance "looks like" conductance (1/resistance), because their product is current. It is this "looking like" that is the basis of the analogy.

> **Warning!** Some authors deal with the analogy the other way around: Φ is said to be analogous to current, and I_{tot} (called "magnetomotive force") is analogous to voltage, in which case reluctance is analogous to resistance rather than conductance. Clearly, this affects the way schematics using the analogy are drawn; but since it is just an analogy, both ways are acceptable, as long as you don't mix them!

In reality, engineers almost never use this analogy, but just stick with equivalent circuit models. An example of the usage of this analogy is given below in the discussion on leakage inductance.

THE IDEAL TRANSFORMER

Let's start our thinking about magnetics by dealing with an approximation, the ideal transformer. It's really a pretty good approximation for many purposes, and understanding it allows us to refine the model later to include nonidealities. A transformer is by definition a magnetic structure that *transforms*: whatever power goes in is what comes out, with no time delay. This is what distinguishes a transformer from an inductor, which can store energy for some time before releasing it.

The ideal transformer has two windings (signified by the curly shapes in Figure 5.3), which sit on a single core. The core is here shown as a magnetic material; if it were an air core, the two straight lines wouldn't be shown. Finally, the direction of current flow (or of applied voltage) is shown by dots: it makes no difference whether a dot signifies the start or end of a winding, as long as it is defined the same way for every winding on the transformer. Physically, this means that there are two ways to wind a wire onto a core (Figure 5.4): the start of the winding can go underneath the core or on top of it. When you actually wind the windings, you just pick one way or the other: if the dot means going underneath for the start, just follow that same rule for every additional winding. As long as you are consistent, the two possible end results are electrically equivalent.

Now let's consider Faraday's law for each side of the ideal transformer. For side 1, we have $V_1 = N_1 \times A_1 \times dB_1/dt$, and for side two $V_2 = N_2 \times A_2 \times dB_2/dt$. Now since the two windings are on the same core, they have the same area, $A_1 = A_2$. And since the

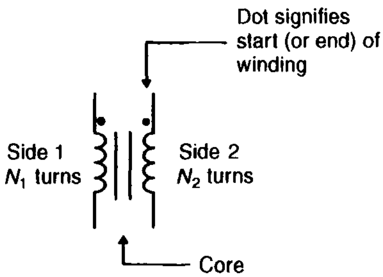


Figure 5.3 In a schematic, a dot signifies the start or end of a winding; its only significance is relative to another dot. The double lines between the windings signify a core.

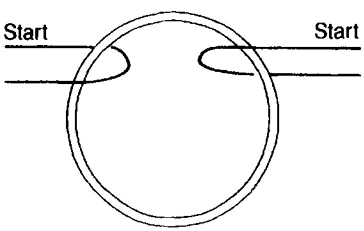


Figure 5.4 Two ways to wind a winding: by starting from the top (right side) or from the bottom (left side).

transformer is ideal, all the flux in one winding is (by definition of what we mean by "ideal") coupled into the other winding, that is, $B_1 = B_2$. Thus

$$\frac{V_1}{N_1} = \frac{V_2}{N_2}$$

which is why a transformer winding is calculated as so many "volts per turn." This equation shows that the volts per turn is the same for every winding on an ideal transformer.

Now an ideal transformer conserves energy, that is, there is no energy storage—exactly whatever is going in is what is coming out, with no time delay. Again, this is the defining property of a transformer. Stated mathematically, this means input power equals output power,

$$V_1 I_1 = V_2 I_2$$

and combining the equations for V, N, and I shows that

$$I_1 N_1 = I_2 N_2$$

that is, if voltage steps up, current steps down.

EXAMPLE

If an (ideal) transformer has 48V in at 2A, and has 24V out, it must have an output current of 4A, because 48V × 2A = 24V × 4A = 96W.

What About a Flyback "Transformer"?

As touched on in Chapter 2, a flyback transformer has the same *name* as a transformer but is physically different. A transformer transforms (power in = power out) an inductor stores energy. A flyback "transformer" acts like both an inductor and a transformer at different times during the switching cycle in a power converter! Perhaps it should have a different name. (Any suggestions? "Transductor" sounds good, since "informer" has another meaning!)

Consider the action of an (isolated) flyback in Figure 5.5 for a moment. When the switch is on, the flyback transformer acts like an inductor. Since the switch is ideally a short circuit when on, a positive voltage is imposed across the primary winding, and so current ramps up in it. Energy is stored in the primary inductance, $E = \frac{1}{2}LI^2$. When the switch is

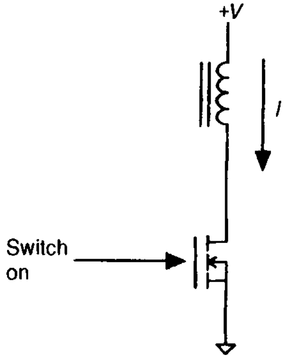


Figure 5.5 When the switch is on, an (isolated) flyback's "transformer" acts like an inductor, storing energy.

off, the flyback transformer acts as a transformer, as shown in Figure 5.6. Since the switch when off is ideally an open circuit, the current has no place to go on the primary, and instead is released on the secondary through the diode. Energy is transferred from the primary to the secondary. The flyback has thus acted as both an inductor and a transformer during a single switching cycle. We'll design a practical flyback transformer later.

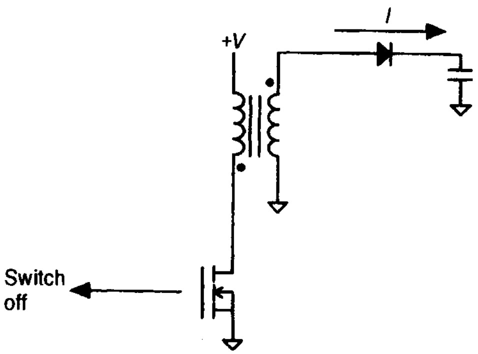


Figure 5.6 When the switch is off, the flyback's "transformer" acts like a transformer, delivering its stored energy to the secondary.

REAL TRANSFORMERS

Real transformers, as opposed to the ideal kind, have many nonidealities. These include nonperfect coupling to the core, core losses, and saturation. Perhaps the most fundamental of these nonidealities is imperfect core coupling; the others are dealt with in subsequent sections.

Nonperfect coupling to the core can come about because of coupling to the air. In Figure 5.7, the magnetic flux in the core set up by one winding doesn't "want" to make the right-angled bend, and a small portion of it escapes into air. Similarly, in a gapped core, the flux is forced to go through a small air gap and some of it does not return to the magnetic material on the other side of the gap, but rather continues out into the air, finding another path for its return. And in a toroid, although the flux is theoretically perfectly coupled to the core, in reality there is always a little bit of nonsymmetry in the winding, and this too causes a tiny bit of coupling to the air. Using the electrical analogy, the circuit of Figure 5.7 could be modeled as shown in Figure 5.8.

The voltage driving the first winding "looks like" a current source in the analogy. The permeability of the core "looks like" a conductance, so it is modeled as a resistor, with a

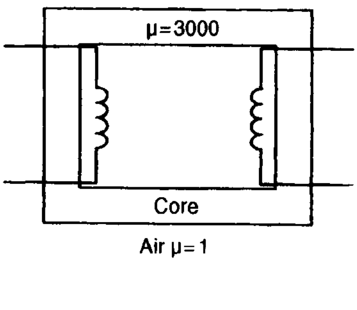


Figure 5.7 Not all of the flux goes through the core, some goes through the air, because the core has finite permeability.

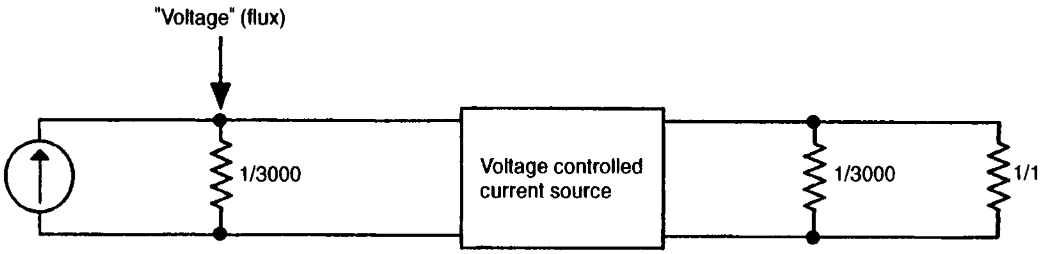


Figure 5.8 The core in Figure 5.7 can be modeled with the electrical analogy.

resistance value that is the inverse of the permeability (1/3000). The resulting voltage (which is what the flux "looks like") is transformed back into a current source on the other side of the core. This reflected current then goes through both the core (resistance = inverse of permeability = 1/3000) and the air [resistance = 1/(the permeability of air), i.e., 1]. The analog of this is that the flux goes through the core and also through the air, with the relative amounts determined by the permeabilities of the two: 3000/3001 of the flux goes through the core, and 1/3001 of the flux goes through the air.

The part of the flux that goes through the air is called the leakage: in the analogy, some part of the current doesn't go through the core "resistor," so the voltage developed across the second winding is smaller than that generated by the primary winding. (The resistors are in parallel, and so the current generates a smaller voltage.) Since some of the flux is not coupled to the secondary, we can now go back and modify our original model of an ideal transformer to take account of this imperfect coupling. In the resulting schematic (Figure 5.9), we still have the perfect transformer in the center of the model. In series with the primary we are showing leakage inductance. The validity of this model is not affected by whether it is shown in the primary or secondary, since it is just subtracting from the voltage that appears on the secondary; here it is shown on the primary side.

EXAMPLE

If the primary has 10 turns and 100μH and 40V and the secondary 20 turns and 400μH and 80V, then the secondary has four times the inductance of the primary (inductance goes as the square of the number of turns). Thus if the primary is shown with a leakage inductance of, say, 1μH, on the secondary, this would appear as 4μH (square of the number of turns.) This makes sense because the

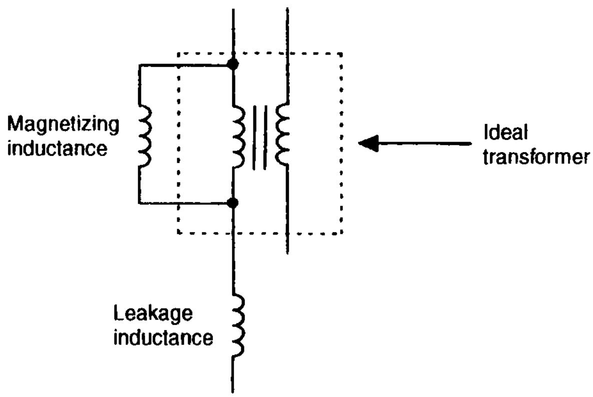


Figure 5.9 A real transformer has magnetizing inductance and leakage inductance, both of which interfere with its ideal transformer action.

leakage inductance causes the same percentage voltage drop whichever side of the transformer it is on: 1μH corresponds on the primary to one turn [(10 turns)^2 = 100μH, so (1 turn)^2 = 1μH], 4μH on the secondary corresponds to two turns [(20 turns)^2 = 400μH, so (2 turns)^2 = 4μH], and on the primary this is 4V/turn × 1 turn = 4V = 10% of 40V, and on the secondary it is also 4V/turn × 2 turns = 8V = 10% of 80V. Labeling Figure 5.9 makes this very clear—try it!

Practical Note Leakage inductance is caused by coupling through the air, not the core. This important fact implies that the amount of leakage inductance for a design depends only on the geometric shape of the coil; *the leakage inductance of a transformer is independent of the material on which the windings are wound.*

Also shown in Figure 5.9 is the magnetizing inductance. Since the core material has finite permeability, the primary has finite inductance (and so, for that matter, has the secondary). This means that applying a voltage to the transformer generates a current, the "magnetizing current" that is merely wasted as far as the transformer action goes (it is not coupled to the secondary). This is why it is shown in parallel with the primary (where the primary is assumed to be the winding with the impressed voltage, not the load). The magnetizing current I_m is determined by

$$V = L_m \frac{dI_m}{dt}$$

Thus, ultimately, both magnetizing inductance and leakage inductance are associated with losses, because they refer to energy that is required to use the transformer but that doesn't end up on the secondary where it can be applied to the load. These inductances are part of what makes the efficiency of a real transformer less than 100%.

Core Materials

Another aspect of a real transformer (or inductor) is its use of real core materials. Not only do real core materials have finite permeability, they have losses, they saturate (what this means is discussed in the next section), and at least some types have permeability, losses, and other properties that are temperature dependent! Incorporating all these factors properly requires some experience and knowledge. Later in this chapter we execute several practical magnetics designs to explore these issues in great detail. For the moment, let's consider Table 5.3, an overview of some practically important types of core materials and some of the pros and cons of using them.

Saturation

The preceding section mentioned "saturation" several times, so I'll explain it right away. Saturation is what happens to a core when it has more than a certain flux density: its permeability is reduced from a high value to approximately 1. This in turn means a radical reduction in inductance, which would clearly be disastrous in some circuits; complete saturation of a core is thus to be avoided in most cases.

TABLE 5.3 Core Materials: Pros, Cons, and Usage

Material	Consideration
Air	*Pro*: Air core magnetics can't saturate! *Con*: The permeability of air is one, so you can only get a small inductance. Practically, this means a couple of microhenrys tops is all you can expect to get from an air core winding. Further, there is of course a lot of fringing, almost by definition! This causes losses, and EMI. *Used*: The primary use of air core magnetics is in rf circuits, where a few microhenrys goes a long way. There has also been occasional talk of applications in ultrahigh frequency power converters, but this could never be practical because of EMI considerations.
Ferrite	*Pro*: Ferrite materials (made by a wide variety of vendors) have high permeability and thus can be used to generate high inductance. The permeability is relatively constant with flux density, and there exist a variety of ferrites optimized for minimal power dissipation in various frequency bands. The poorly controlled initial permeability of ferrites is frequently not a problem, as ferrite cores are often gapped. *Con*: Ferrite saturates hard. *Used*: Ferrites typically are used in power transformers, or for noise filters.
Molyperm (MPP)	*Pro*: MPP cores have a soft saturation. A wide variety of different permeabilities is available, and the permeability can be very well controlled by the manufacturer. *Con*: At typical power supply switching frequencies, MPP cores have much higher losses than ferrites. *Used*: MPP cores are used for inductors or noise filters at high DC currents.
Powdered iron	*Pro, Con*: Powdered iron saturates slightly harder than MPP, and while a variety of permeabilities is available, these are typically lower than what can be had from MPP. This means a powdered iron inductor will be larger than a device having the same inductance and current capacity but built on an MPP core. The big plus is that powdered iron cores are cheaper than molyperm cores. *Used*: The same places that use MPP cores, but where cost is more important than size.
Steel laminations	*Pro*: Steel has a very high saturation flux density, producing very high inductance. Don't ignore this material just because it's old! In some applications, such as very high density converters, steel laminations may be the only way to go. *Con*: For many applications, steel is unaffordably expensive, not to mention heavy. It also saturates hard and has higher losses at high frequency than ferrite. On the other hand, take a look at the new amorphous material, which overcomes some of these limitations. *Used*: Power inductors, low frequency power transformers.

"Saturation" has a fairly clear meaning for ferrites and steel laminations because the core saturates rather abruptly (hard): an extra oersted or two of flux density, and suddenly the permeability plummets (but bear in mind that even for these materials saturation flux density is a function of temperature).

For MPP cores, however, the reduction in permeability as a function of flux is very gradual (soft saturation), and indeed MPP cores are routinely run at reduced permeability intentionally; for this material, the term "saturation" really doesn't have a strict meaning.

> **Practical Note** From a practical standpoint, if current flow has reduced the permeability of the core below, say, 20% of the permeability it has with zero current, the core may be considered to be effectively saturated.

EXAMPLE

The ferrite material 3F3 made by Philips [1] can take > 5000 gauss at 25°C. Applying additional flux (H = oersteds = amp-turns/cm) does not result in much increase in B, as indicated in Table 5.4.

TABLE 5.4 As Flux Increases, Permeability Drops: 3F3 Material

H (Oe)	B (G)
0.5	1000
1.0	2000
1.5	3000
2.0	4000
2.5	5000
3.0	5100

Starting from 0 flux, adding half an oersted at a time increases B by 1000G at a time, until suddenly the core saturates at 5000G: adding an additional half-oersted increases B by only 100G; another half-oersted might add only a few gauss, until you get only one gauss per oersted, that is, a permeability of 1; the core is saturated!

Other Core Limitations

Let's also mention a few of the other nonlinearities of cores that can prove important in practical applications.

Curie Temperature This is the temperature above which the core becomes "demagnetized" and irrecoverably loses its permeability. For 3F3 material the Curie temperature is above 200°C, so typically bobbins would melt or even wire insulation fail before this point was reached. On a toroid, however (since there is no bobbin), it might be possible to reach this temperature using high temperature wire insulation; obviously this is a no-no.

Core Losses Whenever there is a changing flux in a core, there is some power lost, power that goes into heat. The amount of power dissipated is a complicated function of many variables, such as peak-to-peak flux density, temperature, frequency, and of course core material. Observe however, that DC flux density <u>does not</u> cause core loss: a DC inductor with no AC ripple current through it has <u>no</u> core loss!

> **Practical Note** The curves and equations shown in core materials' data books for power loss are for sine waves only! If there is a DC component to the flux, or if the flux is nonsinusoidal, the losses will be different because the B-H loop is different. No one really knows how to calculate the losses for the cases (you can't just decompose the flux into its frequency components as in a Fourier spectrum, because loss is nonlinear in B). Use the sinusoid as an approximation—if you have to know core losses exactly, measurement is the only choice.

Another Limit MPP permeability has a frequency dependence, too.

Optimum Design

It can be shown that minimum power is lost in a magnetic structure when

A. Core losses = copper losses
B. Primary copper loss = secondary copper loss

Practically, these conditions have three implications:

1. For a given magnetic, if the power loss in the core is much less than the copper losses (primary and secondary windings together), you need to decrease the number of turns; this increases flux density and core losses, while decreasing copper losses. The result will be that the total losses decrease.
2. Conversely, for a given magnetic, if the power loss in the core is much greater than the copper losses (primary and secondary together), you need to increase the number of turns; this decreases flux density and core losses, while increasing copper losses. The result will be that the total losses decrease.
3. Allot the same winding area to the primary and secondary; if the secondary has more turns, it must have proportionately smaller wire. If there are multiple secondaries, allot their winding area by output power (i.e., higher output-power outputs get more winding area, so that each secondary has the same I^2R loss).

Naturally, these are only guidelines; in most cases additional considerations (such as being able to get only integer numbers of turns on a winding) will prevent you from following these suggestions exactly. But you don't want losses to be unbalanced by, say, 3 : 1; if they're different by 20–30% that's pretty good.

PRACTICAL DESIGN OF A DC INDUCTOR

Throughout Chapters 5 and 6, much of the design work is targeted at a specific design, a buck converter. Concentrating on this design will help to focus our efforts, because at the end of the design, in Chapter 6, we look at measurements of the converter as built in the lab, and compare them with the results expected from the design work in these two chapters. The measurements will turn out to compare very closely indeed with the designed results.

With all the foregoing background material under our belts, we're ready to design a real DC inductor. A DC inductor is by definition a magnetic structure that has a single winding, invariably on a core, and it carries primarily DC current; that is, the ripple current (AC component of the current) is relatively small.

For our design of a DC inductor, we are going to be working on the choke for the buck converter we are designing. The specific requirements are set by the design: we need 35μH at a DC current of 2A. The power dissipation we've allotted for this component of the design is 300mW, so the maximum resistance should be less than $(2\text{A})^2R = 300\text{mW} \to R = 75\text{m}\Omega$. Observe that at a 250kHz switching frequency, with 15V at the input and 5V on the output (so that duty cycle = 5V/15V = 33%), the ripple current is $dI = V(dt/L) = (15\text{V} - 5\text{V})(33\% \times 4\mu\text{s})/35\mu\text{H} = 0.377\text{A}$ peak-to-peak, which is small relative to 2ADC output, satisfying the requirements for this to be a DC inductor. In addition, since we know that some core types (in particular MPP) used for DC inductors can change permeability with flux, let's require that the core not swing more than 20%. This means that the inductance should not lose more than 20% of its value as the current increases from 0A to 2A; that is, the inductance with no DC current should be 44μH (44μH × 80% = 35μH).

Core Selection

First, let's choose a core material. Since this is a DC inductor, the typical choice would be either MPP or powdered iron. To keep this chapter to a manageable size, we're going to say that getting a small inductor is more important than cheapness, which points to the choice of MPP. In reality, you might try it both ways, to see what makes sense given the overall dimension and cost budget available to the design. Exactly the same procedure detailed in the following sections would be used to design a powdered iron inductor.

First Try

We're going to use a recent Magnetics Inc. catalog [2] to design our inductor, since its selector guide (Figure 5.10) provides a convenient starting point. Without a selector guide, we would have to rely on experience with previous designs for our first guess. As will become apparent, however, making a good first guess is not essential, it merely reduces the amount of work you have to do.

The selector guide (Figure 5.10) bases its guidance on the amount of energy the inductor will have to store (actually twice that amount). The selector guide wants the inductance in millihenrys: 35μH = 0.035mH; at 2A, (twice) the energy is $(2\text{A})^2 \times 0.035\text{mH} = 0.14\text{mJ}$. Tracing along the guide as in Figure 5.11, we find a recommendation of a 200μ core (200 is the initial permeability). In Figure 5.12, we find the core number is 55127.

So our first try is going to be the core 55127. Looking at Figure 5.12, the A_L for the core is 85. What is A_L? It is a convenience provided by the manufacturer, telling you how many millihenrys you get on this core for 1000 turns, or equivalently, how many nanohenrys you get for 1 turn (inductance goes as N^2, so the 1000 : 1 turns ratio means 1,000,000 : 1 for inductance, the same ratio as mH to nH). So to get 35μH, we need:

$$N = \sqrt{\frac{L}{A_L}} = \sqrt{\frac{35\mu\text{H}}{85\text{nH}}} = 20.29 \rightarrow 20 \text{ turns}$$

[*Check*: $(20 \text{ turns})^2 \times 85\text{nH} = 34\mu\text{H}$. The turns are rounded off because of course only an integer number of turns is possible on a toroid winding.]

Now we will calculate the flux density, so we can find out how much the permeability changes between 0 and 2A of DC current. First looking at Figure 5.11, we have that $H/NI = 0.467$. (We can check this, or calculate it for manufacturers who haven't provided it: $H/NI = 0.4\pi/\text{path length} = 0.4\pi/2.69\text{cm} = 0.467$. This is clearly a constant that partially describes the core.) Thus the magnetic field being applied is $H = H/NI \times N \times I = 0.467 \times 20 \text{ turns} \times 2\text{A} = 18.7$ Oe.

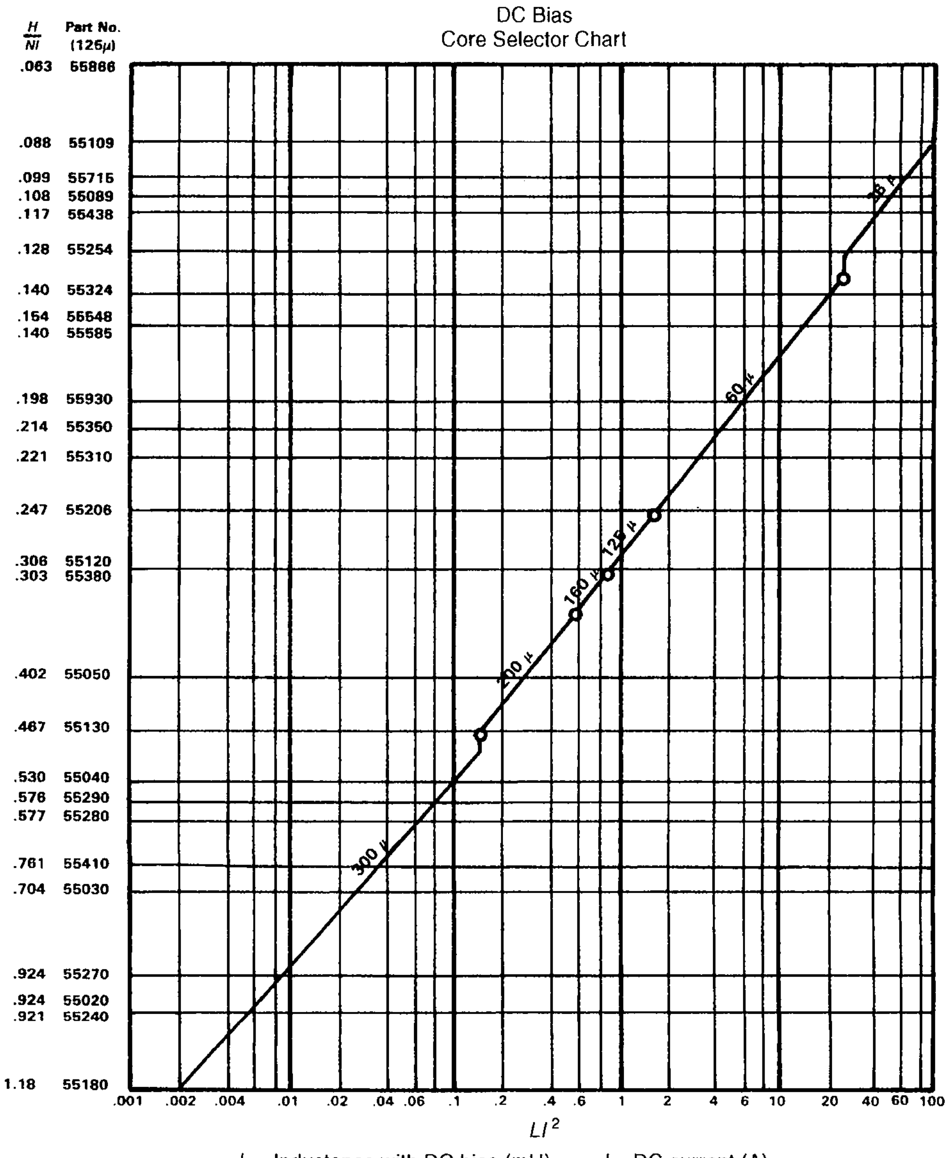


Figure 5.10 A vendor's selector guide (From Ref. 2, p. 31).

We could also calculate the flux density, B, inside the core (it is just the permeability times H), but we will be interested in that only when we get to core losses. For the moment, we need to get the inductance right, and for that we want only the percentage of initial permeability due to the DC current. *Note*: Some manufacturers give percent permeability at just one or two points, making it difficult to know exactly where you are in inductance; I recommend staying away from such materials.

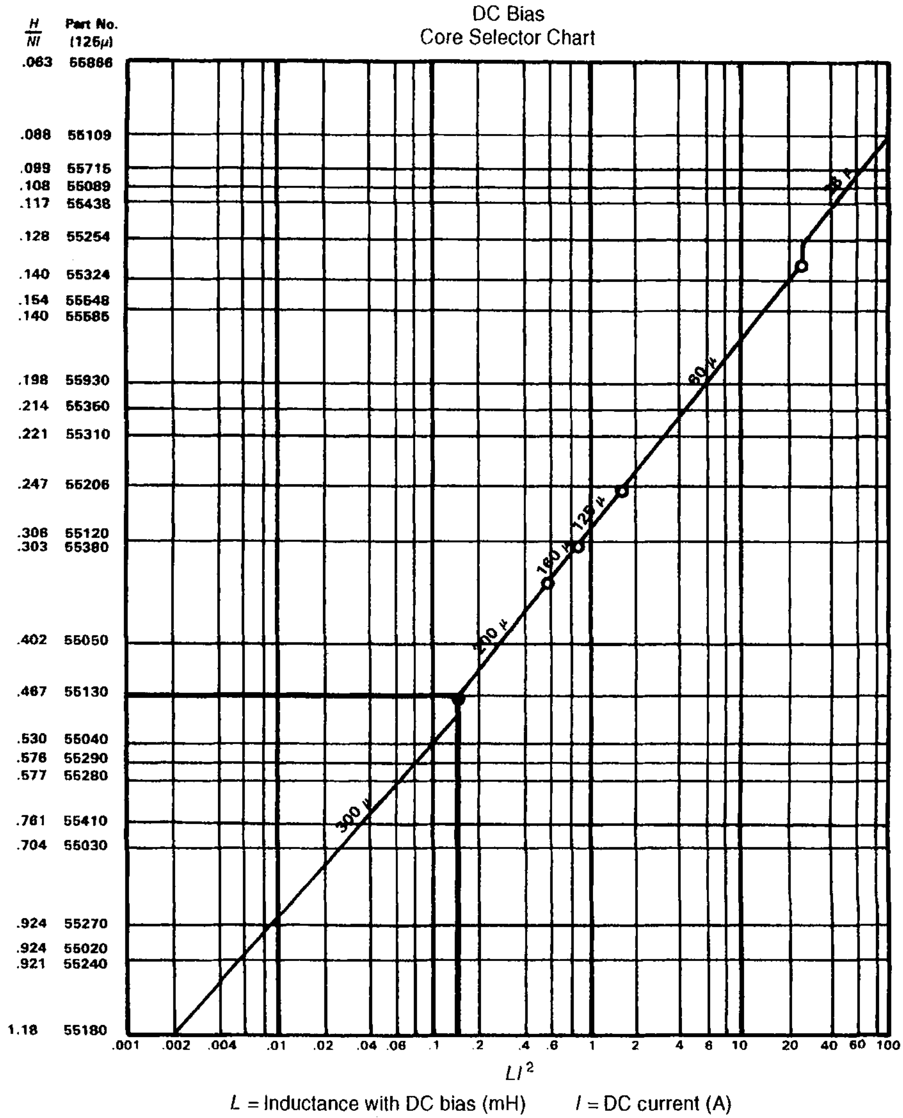
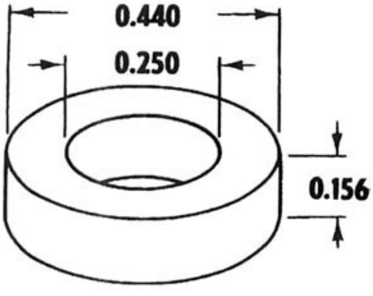


Figure 5.11 Using the vendor's selector guide to make an initial guess for the core. (From Ref. 2.)

> **Practical Note** Manufacturers nowadays also provide equations (as opposed to curves) that describe the permeability as a function of flux; because these equations are fits to data, not based on theory, below about 20% of initial permeability the equations start giving seriously erroneous errors. **Always use the manufacturer's curves, not the equations**. As a further specific note, I observe that the version of the Magnetics MPP catalog I was using had the numbers for the equation for 300μ material with errors in it!



CORE DIMENSIONS AFTER FINISH

OD (MAX.)	0.468 in	11.90 mm
ID (MIN.)	0.232 in	5.89 mm
HT (MAX.)	0.186 in	4.72 mm

WINDOW AREA		53,800 cir. mils
CROSS SECTION	.014 in²	.0906 cm²
PATH LENGTH	1.08 in	2.69 cm
WEIGHT	.0740 oz (.0046 lb.)	2.1 gm

WINDING TURN LENGTH

WINDING FACTOR	LENGTH/TURN	
100% (UNITY)	0.0720 ft	2.195 cm
60%	0.0637 ft	1.942 cm
40%	0.0556 ft	1.695 cm
20%	0.0520 ft	1.585 cm
0%	0.0507 ft	1.545 cm

WOUND COIL DIMENSIONS

UNITY WINDING FACTOR

OD (MAX.)	0.618 in	15.7 mm
HT (MAX.)	0.353 in	9.0 mm

MAGNETIC INFORMATION

PART NO.	PERM μ	INDUCTANCE @1000 TURNS (mH±8%)	NOMINAL DC RESISTANCE (Ohms/mH)**	FINISHES AND STABILIZATIONS*	GRADING STATUS 2% BANDS	B/NI (Gauss/amp-turn)
55133–	14	6	4.10	A2	•	6.54 (<1500 gauss)
55132–	26	11	2.23	A2	•	12.1 (<1500 gauss)
55131–	60	26	.945	ALL	•	28.0 (<1500 gauss)
55130–	125	53	.464	ALL	YES	58.4 (<1500 gauss)
55129–	147	63	.390	ALL	YES	68.6 (<1500 gauss)
55128–	160	68	.361	ALL	YES	74.7 (<1500 gauss)
55124–	173	74	.332	ALL	YES	80.8 (<1500 gauss)
55127–	200	85	.289	ALL	YES	93.4 (< 600 gauss)
55125–	300	127	.194	A2 & L6	YES	140 (< 300 gauss)

14 through 160μ types are available as high flux cores.

WINDING INFORMATION

FOR UNITY WINDING FACTOR**

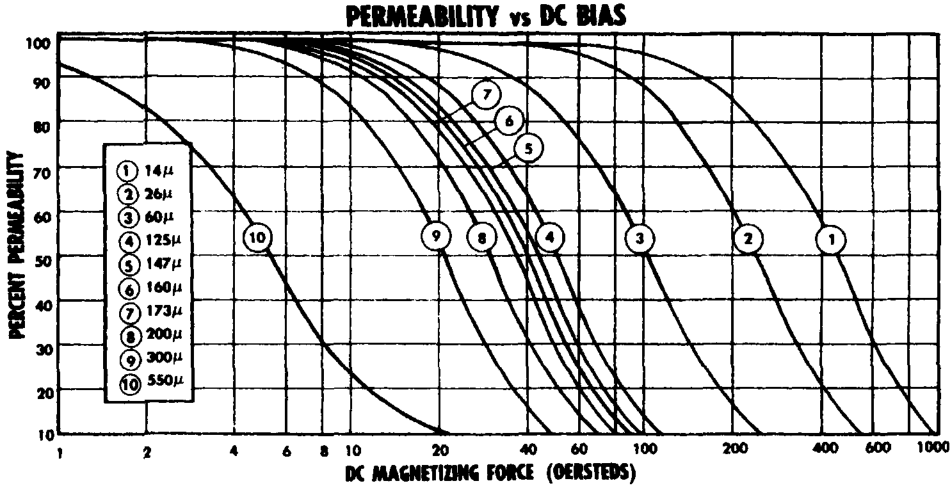
AWG WIRE SIZE	TURNS	R_{DC} (Ohms)	AWG WIRE SIZE	TURNS	R_{DC} (Ohms)
17	22	0.00808	31	460	4.34
18	28	0.01276	32	560	6.54
19	34	0.02000	33	695	10.31
20	43	0.03192	34	885	16.63
21	54	0.04958	35	1,098	26.2
22	66	0.07747	36	1,355	40.5
23	83	0.1210	37	1,655	61.0
24	102	0.1896	38	2,069	96.5
25	127	0.295	39	2,663	162.4
26	158	0.467	40	3,363	259
27	199	0.737	41	4,138	393
28	245	1.150	42	5,275	630
29	299	1.747	43	6,405	987
30	374	2.80	44	7,370	1,374

*Further stabilization and grading information available from the vendor.
**The nominal DC resistance and the R_{DC} are theoretical values not attainable in practice.

Figure 5.12 Vendor's data sheet that applies to the core selected, number 55127. (From Ref. 2, p. 54.)

From this, we can find the percent of initial permeability from a curve the manufacturer provides (Figure 5.13).

The 55127 core is 200μ, so referring to Figure 5.13, we are going to use curve 8. With 19 Oe, the percentage of initial permeability of this core is 75%. This means that the inductance at 2A is reduced to only 75% × 34μH = 25.5μH. To increase this inductance, we would have to add turns—but we are already past the 80% swing point. Adding turns would



General equation:

$$\mu_e = \frac{\mu_i}{100}\sqrt{\frac{a+cH+eH^2}{1+bH+dH^2}}$$

where H = DC bias in oersteds

	μ_i	a	b	c	d	e
High Flux	14	10004	.0027514	26.879	.000011888	.041784
	26	9914.0	-.000088616	2.4260	.000037542	.012877
	60	10159	.00058116	-41.642	.000047498	.050984
	125	9952.5	-.0046907	-59.727	.00041074	.11038
	160	9993.9	.0017796	-107.06	.00047937	.29598
MPP	14	9949.6	-.0018420	-22.199	.0000045774	.013253
	26	9979.7	-.0039552	-42.541	.000018943	.050104
	60	10067	-.0054844	-63.702	.00012792	.10098
(Graph above)	125	10174	-.015802	-169.63	.00051688	.76876
	147	10279	-.016889	-221.49	.00054472	1.2934
	160	10238	-.018887	-220.62	.00074733	1.2676
	173	9996.2	-.017441	-157.84	.0012421	.34560
	200	10025	-.027999	-304.90	.0012754	2.5101
	300	10021	-.034600	-386.56	.0011600	4.2854
	550	9956.8	-.016354	-1054.6	.044933	34.653

Figure 5.13 Determining the percentage of initial permeability of the design according to the vendor's data. (From Ref. 2, p. 29; references to material not included in this book deleted.)

increase the flux density, even further increasing the swing. So instead, let's try a core with a lower permeability.

Second Try

For a second try, let's use a 125μ core (60, 125 and 300μ seem to be the easiest perms to get hold of). In this core size, this is the 55130 core. This core has an $A_L = 53\text{nH}$, so we need:

$$N = \sqrt{\frac{L}{A_L}} = \sqrt{\frac{35\mu\text{H}}{53\text{nH}}} = 25.7 \rightarrow 26 \text{ turns}$$

Again calculating, $H = 0.467 \times 26$ turns $\times$ 2A = 24 Oe. This is higher than before, but remember that this is a lower perm core, and therefore a higher flux density doesn't necessarily mean a lower percent permeability!

Looking at Figure 5.13 again, a 125μ core is curve 4. At 24 Oe we have 80% of initial permeability. Now the actual inductance achieved is $L = A_L \times N^2 \times \%\text{perm} = 53\text{nH} \times 26^2 \times 80\% = 28.7\mu\text{H}$. We need to get this up to 35μH, so remembering that inductance goes as turns squared, we need to increase the turns to:

$$N = \sqrt{\frac{\text{desired inductance}}{\text{actual inductance}}} = \sqrt{\frac{35\mu\text{H}}{28.7\mu\text{H}}} = 29 \text{ turns}$$

We have $H = 0.467 \times 29$ turns $\times$ 2A = 27 Oe, which is still 80% permeability, and, at last, $L = 53\text{nH} \times 29^2 \times 80\% = 35.7\mu\text{H}$. As a sidelight, $B = H \times \mu \times \%\text{perm} = 27$ Oe $\times$ 125 $\times$ 80% = 2700G. Remember that this is DC flux density, and thus does not contribute to losses.

Practical Note The most common cores seem to be 60, 125 and 300μ. If you're in a rush to get a sample, you may find it best to select one of these.

Let's suppose that after a couple iterations like this (they take only a minute each), we find that the lowest perm core in this size doesn't make it. Or suppose the 60μ core doesn't make it, and we don't want unusual core types. Or again, suppose we didn't follow the selector guide and need to get the absolutely smallest possible core for the application. We then have three choices.

1. We could relax the criterion we had pre-established concerning the maximum swing in the core, allowing the inductance to start at a higher value. The effect of this is that the ripple current in the inductor, and thus the ripple voltage on the output capacitor, changes more from minimum to maximum load—perhaps not a big deal. It also means that the double-pole frequency set by the inductor and the output capacitor will move more—but again, perhaps this can be lived with.
2. We could go to a *high-perm* core, which are those (for Magnetics, Inc.) that start with a 58XXX rather than a 55XXX. These cores take a higher flux before saturating. Since, however, most of the improvement is in the range of 50% initial perm and up, this probably isn't going to help unless you've already implemented suggestion 1. In addition, all the high perm cores are special order, and cost somewhat more.
3. The final choice, obviously, is to go to the next core size up, and try again.

All these options take only a few minutes to explore, so a core can usually be selected within a quarter-hour.

Selecting the Wire

Now that we have selected a core and the number of turns to go on it to get the inductance specified, we can calculate the wire size that fits on this core, and thence the power

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dissipated. The process of wire selection is similar for all magnetic structures, so we're going to do it in great detail here, refering you back to this section when it's time to select wire for other magnetics designs.

From the data book (our Figure 5.12), the winding area of the 55127 core is 53,800 circular mils. (You were *warned* that these crazy units show up sometimes!) Now for a toroid, you can't actually fill up this entire winding area, because you couldn't get a winding tool into the tiny hole that's left when the core is wound. Besides, wire doesn't lay (pack) neatly. In practice, then, the best filling you can get on a toroid is 45–50% of the winding area (this is called the *fill factor*: see Figure 5.14).

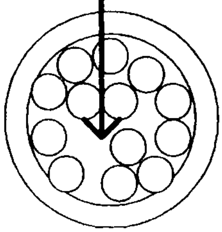

Figure 5.14 Wire can't fill entirely the winding area; 50% is typical maximum fill factor.

Don't forget, you also have to include the cross-sectional area contributed by the wire insulation! There's double (heavy), triple, and quad insulation, and all have different areas. It's surprisingly common for a designer to just pull out a table for bare copper wire, make a selection without thinking about it, and end up not being able to get those last couple of turns on. Also, small wire has its insulation as a greater percentage of its total cross section, and with the multiple thin wires known as litz, insulation can take up 50% of the available winding area!

So the cross-sectional area available for a single turn is going to be half the total winding area divided by the total number of turns:

$$A_L = \frac{\text{winding area fill factor}}{\text{number of turns}}$$

$$= \frac{53{,}800\text{c.m.}/0.5}{29} = 928 \text{ c.m./turn}$$

Referring to Figure 5.15, we select the closest size, 22AWG, rounding down in area to avoid exceeding our 50% fill factor.

Calculating the Resistance

Having selected the wire gauge, we can now calculate the resistance of this winding on this core. Referring again to Figure 5.12, we are going to use the "100% fill factor" length per turn number. This choice has both a practical and a theoretical justification. Theoretically, our 50% fill factor is going to just about fill up the core, again because of insulation, packing, etc. Thus the 100% number is closer to reality. Practically, it has been found by winding many different cores over the years that the 100% gives a better approximation—and it's usually better to have an overestimate of the winding resistance than an underestimate.