

Power Supply Design Seminar

Practical Magnetic Design: Inductors and Coupled Inductors

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Practical Magnetic Design: Inductors and Coupled Inductors

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ABSTRACT

Magnetics design has always been somewhat of a mystery to many analog design engineers. In this paper, I will attempt to demystify this by first discussing the foundational equations and conversion of units necessary for magnetics design. I will give an in-depth example for a flyback-coupled inductor, showing how to calculate inductance, wire gauge and AC/DC losses, flux density, core and bobbin fill factor, and inductance rolloff. I will then discuss various core materials and how to get good coupling by properly layering the primary and secondary windings on a bobbin.

I. INTRODUCTION

Switch-mode power supply (SMPS) engineers must have a working knowledge of magnetics design and function in order to truly understand switch-mode power conversion. Many books and papers have been written on the theory and function of magnetics, but there is sometimes a disconnect between theory and foundational concepts and the actual application of that theory to build a magnetic component.

In this paper, I'll cover the necessary equations and calculations you will need when designing a magnetic inductor or coupled inductor. A flyback magnetic is also known as a coupled inductor because it stores energy in one half-cycle and then delivers the energy to the secondary on the next half-cycle, whereas transformers receive and deliver energy to the secondary within the same cycle.

A step-by-step approach will include enough theory to support the concepts and equations provided.

II. DESIGN FLOW DIAGRAM

Figure 1 shows the design flow diagram for a coupled inductor. Before starting the design, you must obtain a specification that includes inductance, turns ratio, peak and RMS currents, output power, and frequency. You can select a core based on output power, but there are a number of trade-offs you'll need to make to determine which core to use for a given application. (For more

information, see section 4 of "Power Transformer Design" by Lloyd H. Dixon.) The bottom line is that there is no way to calculate the right core for a given application; there are too many trade-offs involved. The best you can hope for is to get close.

When it comes to core selection, there is no replacement for experience. The experience gained from iterating through a design many times and building and testing multiple designs to see which works best is invaluable. I have found by experience that economical flat design (EFD) cores allow easier interleaving of the secondary winding between two halves of the primary because the winding window is wide – I'll address this topic more extensively later on.

The core and bobbin combinations that lend themselves to proper layering of the windings on the bobbin yield better coupling and lower leakage. These are two very important parameters when designing a flyback magnetic.

Once you have the core inductance, you can calculate wire losses and flux density. If the turns don't fit evenly on the bobbin, you may need to change the core gap, which in turn changes the Al , to get the right amount of turns on the bobbin to fit evenly. If the flux density is too high, you may need a bigger core or different core material.

After determining the right core and material, you can then calculate the bobbin factor. If the bobbin factor is too high or the turns don't lie properly on the bobbin, the gap of the core can be changed so that the bobbin will take more or less turns to make the same inductance.

The wire gauge can also be changed to fit on the bobbin properly. That is why this is an iterative process; there are a number of ways to solve the same problem. Only designers can determine which solution is best for a given application.

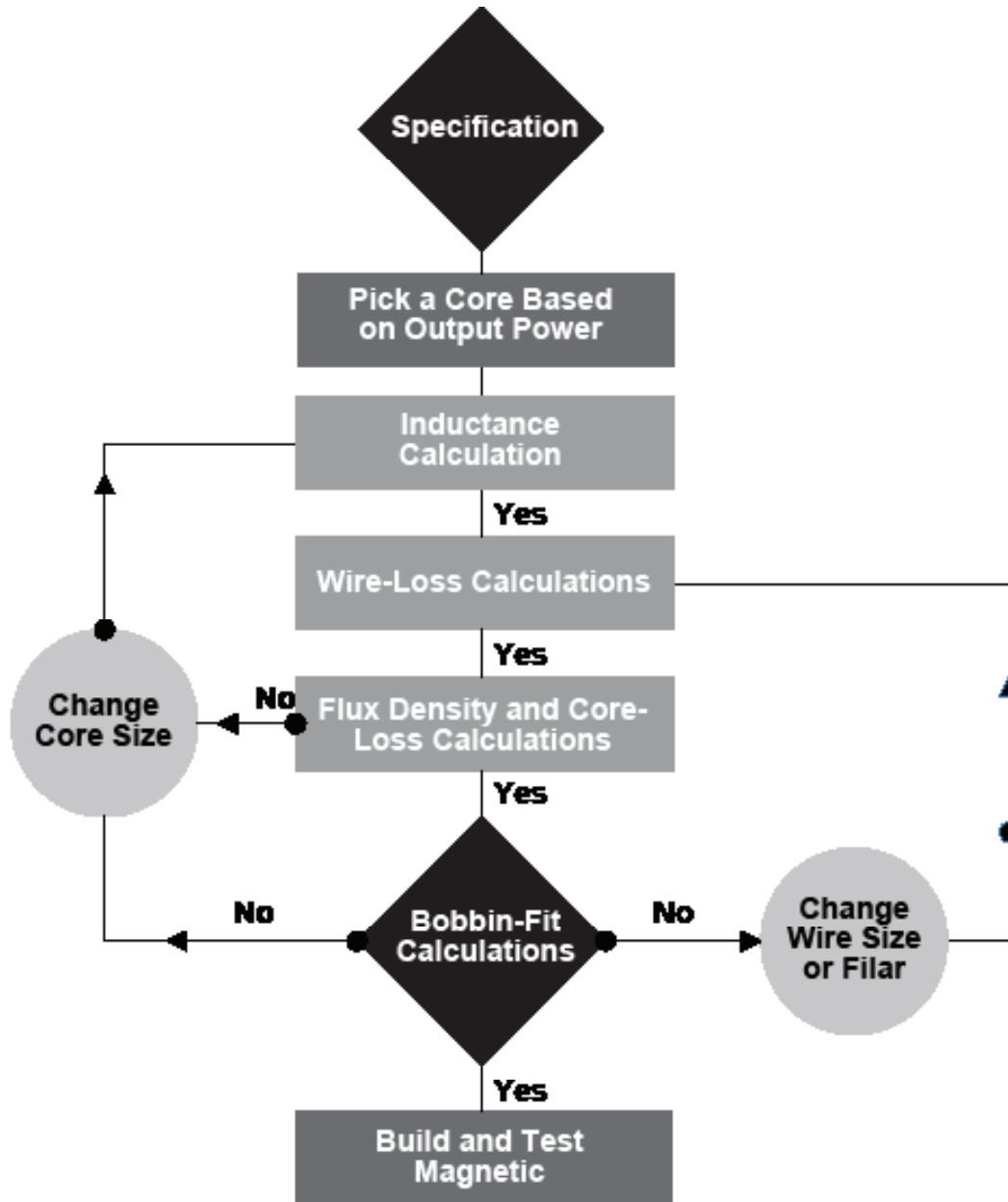


Figure 1 – Design flow diagram.

III. MAGNETIC PARAMETERS AND CONVERSION FACTORS

The International System of Units (SI system) is an internationally accepted system in which permeability is rationalized to free space, which is $4\pi \times 10^{-7}$. This simplifies equations in the SI system. In the centimeter-gram-second (CGS) system, permeability is not rationalized; it

therefore becomes part of the equations in CGS systems and complicates them. You cannot control what system magnetic manufacturers use in their data sheets, so it is often necessary to convert from one system to the other. Table 1 shows the conversion factors for both SI and CGS systems.

		SI	CGS	CGS to SI
Flux Density	B	Tesla	Gauss	10^{-4}
Field Intensity	H	A-T/m	Oersteds	$1,000/4\pi$
Permeability	μ	$4\pi \times 10^{-7}$	1	$4\pi \times 10^{-7}$
Area	Ae	m ²	cm ²	10^{-4}
Length	le, lg	m	cm	10^{-2}

Table 1 – Magnetic parameters and conversion factors.

IV. Inductance

The inductance of a wound core can be calculated by the core geometry using Equation 1. This equation is not in the SI or CGS systems. It was derived using millimeters for dimensions, which is what was given in the Ferroxcube data sheet.

$$\text{Linductance} = \frac{.4 \times \pi \times N^2 \times (\text{Ae}) \times 10^{-9}}{lg + \frac{le}{\mu}} \quad (1)$$

Where:

Linductance = Henry's

μ = core permeability

N = number of turns

Ae = core cross-section (mm²)

le = core magnetic path length (mm)

lg = gap (mm)

Manufacturers usually combine the constants for a given core into a term called *Al*. The inductance can then be calculated directly from Equation 2.

$$\text{Lind} = Al \times N^2 \quad (2)$$

V. CORE GEOMETRY

You can calculate the mean magnetic path length, *le* and core cross-sectional area, *Ae*, for any core by using the dimensions of the core geometry. For example, when using a toroidal core (Figure 2), the magnetic path length can be calculated using Equation 3.

$$le = \frac{\pi(\text{OD} - \text{ID})}{\ln\left(\frac{\text{OD}}{\text{ID}}\right)} \quad (3)$$

Where:

OD = outside diameter of core (mm)

ID = inside diameter of core (mm)

N = number of turns

Ae = core cross-section (effective area) mm²

le = Mean magnetic path length (mm)

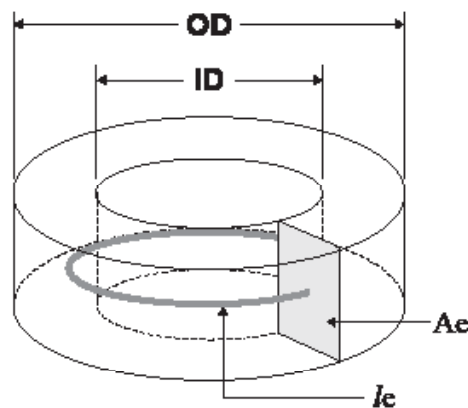


Figure 2 – *le* is a simple calculation with a toroid.

Topic 4

VI. AMPERE'S LAW

Ampere's law states that the total magnetic force along a closed path is proportional to the ampere-turns in a winding that the path passes through. In the SI system, magnetic force is expressed as ampere-turns, A-T, as shown in Equation 4:

$$\begin{aligned} \text{SI: } F &= \oint H \times dl = N \times I \approx H \times le & (4) \\ H &\approx \frac{N \times I}{le} \text{ amper-turns/meter (A-T/m)} \end{aligned}$$

Where:

- H = magnetizing force (ampere-turns/meter)
- N = number of turns
- le = core magnetic path length (m)
- I = peak magnetizing current (amperes)

The equation for magnetizing force in the CGS system is shown in Equation 5:

$$\begin{aligned} \text{CGS: } F &= \oint H \times dl = 0.4 \times \pi \times N \times I \approx H \times le \\ H &\approx \frac{0.4 \times \pi \times N \times I}{le} \text{ (Oersteds)} & (5) \end{aligned}$$

Where:

- H = magnetizing force (Oersteds)
- N = number of turns
- le = core magnetic path length (m)
- I = peak magnetizing current (amperes)

VII. FARADAY'S LAW

The total magnetic flux, Φ , passing through a surface of area, A_e , is related to the flux density, β . The flux rate of change is proportional to the volts per turn applied to a winding. If a secondary winding is coupled to all of the flux in a primary winding, then all of the volts per turn in the primary will be induced in the secondary winding. Lines of flux follow a closed path and have no beginning or end. In the SI system, flux density is expressed in tesla, as shown in Equation 6:

$$\begin{aligned} \text{SI: } \Delta\Phi &= \frac{1}{N} \int E \times dt & (6) \\ E &= N \times \frac{d\Phi}{dt} \approx N \times A_e \times \frac{d\beta}{dt} \Rightarrow \beta = \frac{\int E \times dt}{N \times A_e} \end{aligned}$$

Where:

- β = flux density (tesla)
- N = number of turns
- A_e = effective core area (m²)
- E = voltage across coil (volts)

The equation for flux density in the CGS system is shown in Equation 7:

$$\begin{aligned} \text{CGS: } \Delta\Phi &= \frac{10^8}{N} \int E \times dt & (7) \\ E &= \frac{N}{10^8} \times \frac{d\Phi}{dt} \approx \frac{N \times A_e \times d\beta}{10^8 \times dt} \Rightarrow \beta = \frac{\int E \times dt \times 10^8}{N \times A_e} \end{aligned}$$

Where:

- β = flux density (Gauss)
- N = number of turns
- A_e = effective core area (cm²)
- E = voltage across coil (volts)

VIII. BH CURVES

Figure 3 shows a typical BH curve for a magnetic core. The shaded portion is the BH curve for a converter running in continuous current mode superimposed on the core BH curve. H is proportional to I (current); therefore, as the peak-to-peak current in a magnetic increases, the excursion of the flux also increases.

At the top part of the BH curve, B flattens out; at this point the magnetic is in saturation. Once the current in a magnetic drives the flux to saturation, the core no longer exhibits magnetic properties and the magnetic becomes a wire. Thus it is important to calculate the flux density for a given design to make sure that you don't saturate the core. β_{ac} is peak-to-peak flux density; β_{max} is the peak flux density. β_{max} assumes square wave excitation.

Equation 8 gives the AC flux density, β_{ac} , which is proportional to the peak-to-peak current in a given winding. Equation 9 gives the maximum flux density.

$$\beta_{ac} = \frac{V_{in} \times T_{on}}{A_e \times N_p} \times 10^8 \text{ (Gauss)} \quad (8)$$

$$\beta_{max} = \frac{L_p \times I_p}{A_e \times N_p} \times 10^8 \text{ (Gauss)} \quad (9)$$

Where:

β = flux density (Gauss)

N_p = number of primary turns

T_{on} = on time (sec)

A_e = effective core area (cm²)

V_{in} = voltage (volts)

L_p = primary inductance (Henry's)

I_p = peak primary current (A)

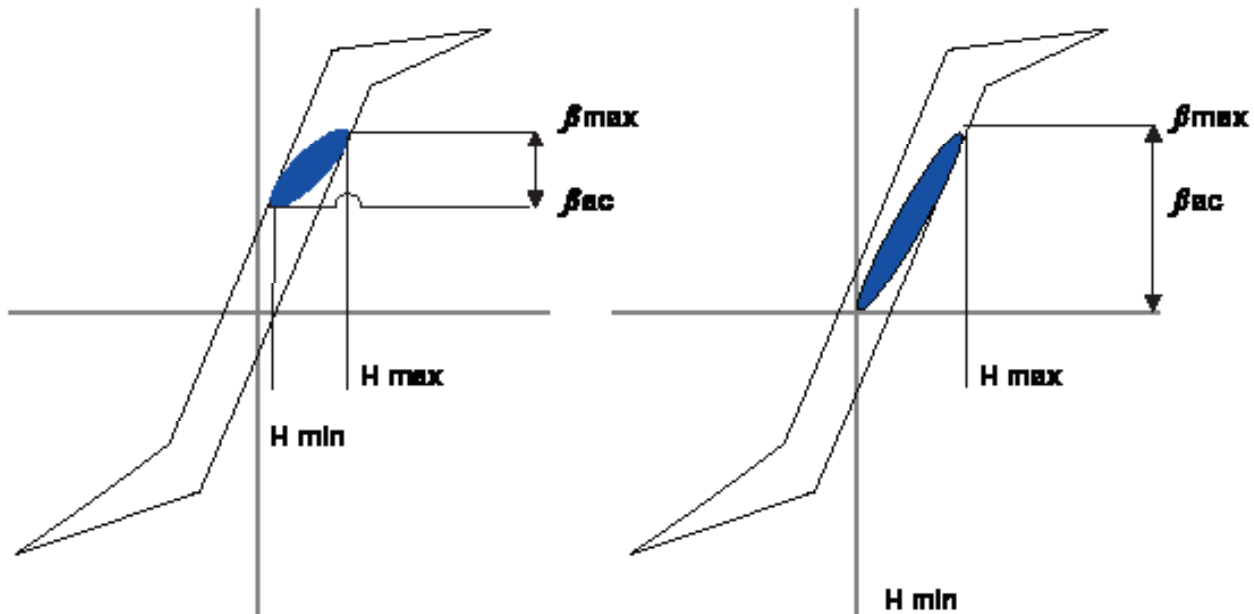


Figure 3 – A typical BH continuous curve (left) and discontinuous curve (right).

IX. CORE LOSS

The power loss in a core is related to half the AC flux density of the core and the frequency applied to the core. To determine core loss, manufacturers normally provide a graph of power loss as a function of peak flux density and core material. Figure 4 shows a graph from Ferroxcube to determine core loss.

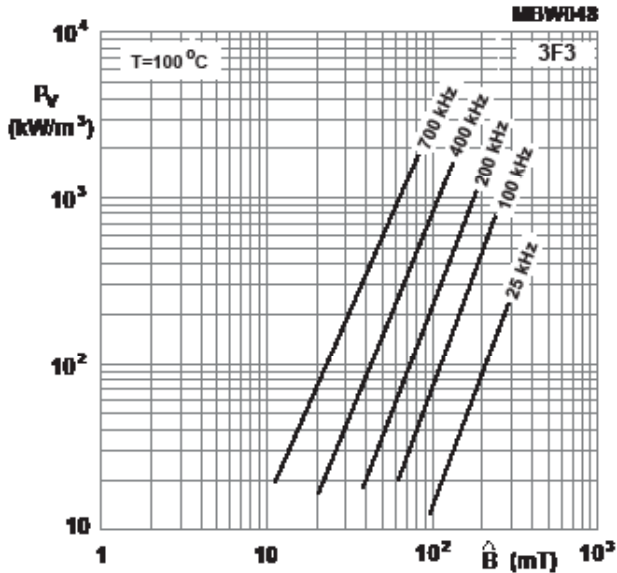


Figure 4 – Power loss as a function of one-half of peak-to-peak flux density (courtesy of Ferroxcube).

To determine the core loss from Figure 3, use Equation 10 to determine the peak flux density in Gauss. (I use Gauss here, to show how to work through the calculation if the manufacturer does not give data in Tesla.) The flux density therefore must be converted to Teslas from Gauss using Equation 11.

$$\beta_{acpeak} \text{ (Gauss)} = \frac{\beta_{ac}}{2} \quad (10)$$

$$\beta_{mt} \text{ (milli Tesla)} = \frac{\beta_{acpeak}}{10} \quad (11)$$

Core loss in Figure 4 is given in KW/m³; the effective volume of the core listed in the data sheet must be used to cancel the m³ term and determine the core loss in terms of watts. The effective volume in the data sheet is given in terms of mm³, so the effective volume, Ve, must be converted from mm³ to m³ using Equation 12:

$$Ve(m^3) = \frac{Ve \text{ (mm}^3\text{)}}{1,000^3} \quad (12)$$

Core loss is given by Equation 13:

$$P_{core \text{ loss}} = Pv \left(\frac{KW}{m^3} \right) \times Ve \text{ (m}^3\text{)} \quad (13)$$

X. PERMEABILITY AND INDUCTANCE ROLLOFF

As current is increased in a magnetic and the flux density gets closer to core saturation, the permeability starts to roll off. Permeability is an important parameter, because as permeability rolls off, the inductance decreases (as shown in Equation 1). If the inductance decreases, the peak-to-peak current goes up in the magnetic; this causes the flux density to increase and push the core closer to saturation. Therefore, permeability can cause a snowball effect as the core gets closer to saturation.

Figure 5 is a graph of permeability as a function of magnetizing force, H, for a given core using 3F3 material.

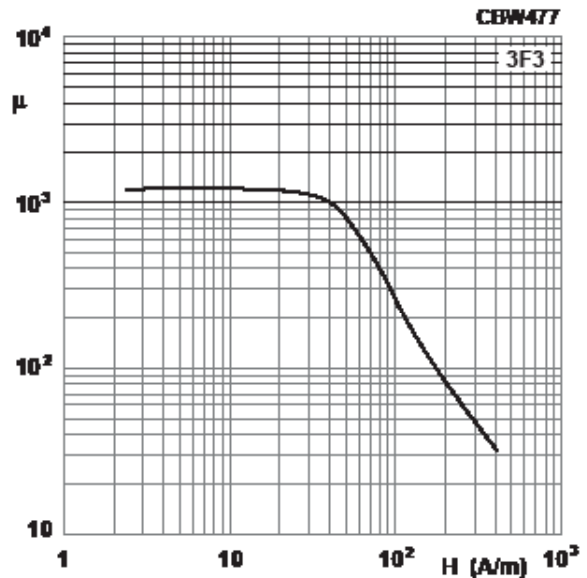


Figure 5 – Permeability as a function of magnetizing force (H) (courtesy of Ferroxcube).

XI. AC/DC WIRE LOSS

Current density, J , is a rule of thumb that says a wire should not handle any more than 400 A/cm². The wire area needed for a specific application is calculated by dividing the RMS current through the wire by 400 A/cm², as shown in Equation 14. This calculation takes into account only the DC portion of the wire loss. You will need calculations for a range of wire gauges because the correct wire gauge has not yet been determined.

$$\text{Wire area required (cm}^2\text{)} = I_{\text{rms}}/J \quad (14)$$

The AC losses are attributed to skin effect. The result is that as frequency increases, current density increases at the conductor surface and decreases toward zero at the center, as shown in Figure 6. The current decreases exponentially within the conductor, and the portion of the conductor that is actually carrying current is reduced, so the resistance and losses will be greater at high frequency than at low frequency.

Penetration or skin depth, D_{pen} , is defined as the distance from the surface to where the current

density is 1/e times the surface current density. Skin depth is given in Equation 15:

$$\delta(\text{skin depth cm}) = \sqrt{\frac{\rho}{\pi \times \mu \times f}} \approx \frac{7.6}{\sqrt{f}} \text{ at } 100^\circ\text{C} \quad (15)$$

Where:

ρ resistivity of copper = $2.3 \times 10^{-6} \Omega\text{cm}$ at 100°C

f = frequency (Hz)

$\mu = 4\pi \times 10^{-7}$

Using look-up tables from reference [3], determine the wire area in cm² and the radius in cm of the chosen wire.

To determine the actual AC resistance of a chosen wire, use Equation 16 once you know the skin depth and the annular inner ring. The annular inner ring is the area of the wire used for AC conductance.

Keep in mind that when the skin depth is greater than the radius of the wire, you will need to set the ring area equal to the area of the wire. This is because the eddy currents have fully penetrated the wire, so smaller wires allow better penetration of the AC current.

$$\begin{aligned} \text{Annular inner ring (cm}^2\text{)} & \quad (16) \\ & = \pi \times \left[\text{radius of wire}^2 - (\text{wire radius (cm)} - \text{skin depth (cm)})^2 \right] \end{aligned}$$

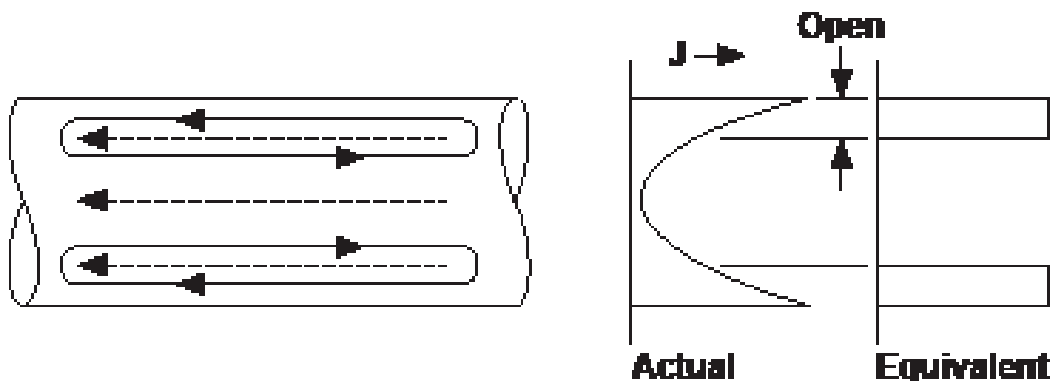


Figure 6 – Skin depth.

Once you've determined the annular inner ring, you can then determine the ratio of AC-to-DC resistance with Equation 17:

$$\begin{aligned} \text{Ratio Rac to Rdc (Rskin)} \\ = \frac{\text{Area of wire (cm}^2\text{)}}{\text{Annular inner ring (cm}^2\text{)}} \end{aligned} \quad (17)$$

Use Equation 18 to calculate the AC and DC wire resistance:

$$\begin{aligned} \text{AC-DC wire resistance} = \\ \text{DC resistance} \times \text{Ratio Rac to Rdc (Rskin)} \end{aligned} \quad (18)$$

When designing a magnetic, it is sometimes desirable to use multiple wires in parallel. You may need to fill a complete layer on the bobbin for better coupling, for example. Another reason to parallel wires is that you can use a diameter less than two times the skin depth and achieve the required area.

Equation 19 shows the relationship between wire area, chosen wire area and the ratio of AC-to-DC wire losses. From Equation 19, you can determine the number of parallel wires.

$$\begin{aligned} \text{Number of wires in parallel} = \\ \frac{\text{Wire area required}}{\frac{\text{Area of chosen wire}}{\text{Ratio of AC to DC losses}}} \end{aligned} \quad (19)$$

Equation 20 is used for calculating the total resistance for a given winding, taking into account AC and DC losses. You can obtain the length per turn for a chosen bobbin from manufacturer data.

$$\begin{aligned} \text{Total winding resistance} = \\ \text{AC-DC wire resistance} \times \text{number of turns} \\ \times \frac{\text{Length per turn}}{\text{Number of wires used}} \end{aligned} \quad (20)$$

XII. PROXIMITY EFFECTS

AC current in a conductor induces eddy currents in adjacent conductors by a process called the "proximity effect." Figure 7 shows a multiwinding transformer, along with a low-frequency total field, F , and energy density diagrams. The field builds uniformly within each conductor according to Faraday's law, and the energy density in the field rises with the square of the field strength. Multiple layers cause the field to build at high frequencies, and the eddy current losses rise exponentially as the number of layers increases.

By interleaving the windings, as shown in Figure 8, the field is reduced and the energy under the energy density curve is reduced to one-quarter of what it was.

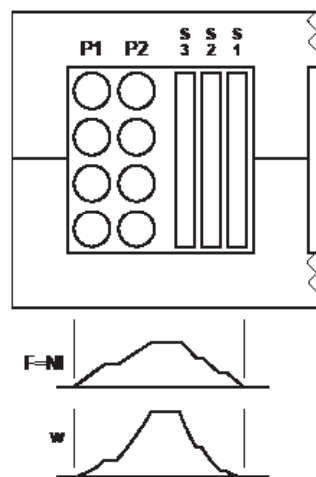


Figure 7 – Multilayer winding.

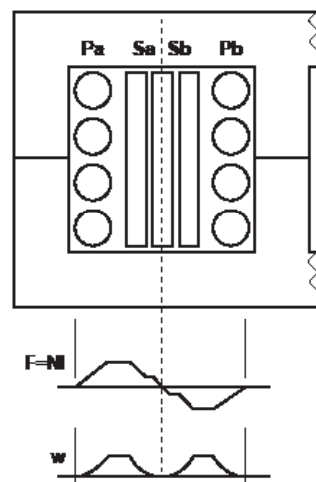


Figure 8 – Interleaved windings.

You can use Equation 21 to determine power loss in a given layer. The design example in this paper does not require taking proximity effects into account because the windings are interleaved and there is only one layer of secondary winding, so proximity effects will not be discussed further.

$$\text{Power loss per layer (m)} = I^2 \left[(m-1)^2 + m^2 \right] \times \left(\frac{\text{Wire diameter (cm)}}{\delta} \times R_{dc} \right) \quad (21)$$

Where:

- m = layer
- I = RMS current
- δ = skin depth (cm)
- R_{dc} = DC resistance of the chosen wire

XIII. WINDING FACTOR AND BOBBIN-FIT CALCULATIONS

One criteria for choosing a particular bobbin is that the windings lay on the bobbin such that very good coupling can be achieved, as discussed previously. The second determining factor is whether the windings fit on the bobbin. In order to determine if the amount of turns needed for a particular magnetic will fit on the bobbin, you must calculate the bobbin fill factor. You can determine the turns per layer using Equation 22 and manufacturer data from Figure 9. In Equation 22, the (-2) is added to save some space for tape.

$$\text{Turns per layer} = \frac{\text{Bobbin width (mm)}}{\text{Diameter of wire (mm)}} - 2 \quad (22)$$

To calculate the number of layers that can fit on a chosen bobbin for a particular wire diameter, first determine the height or buildup available on the bobbin and divide that by the wire diameter. See Equations 23 and 24:

$$\text{Buildup} = \frac{\text{Winding area (mm}^2\text{)}}{\text{Winding width (mm)}} \quad (23)$$

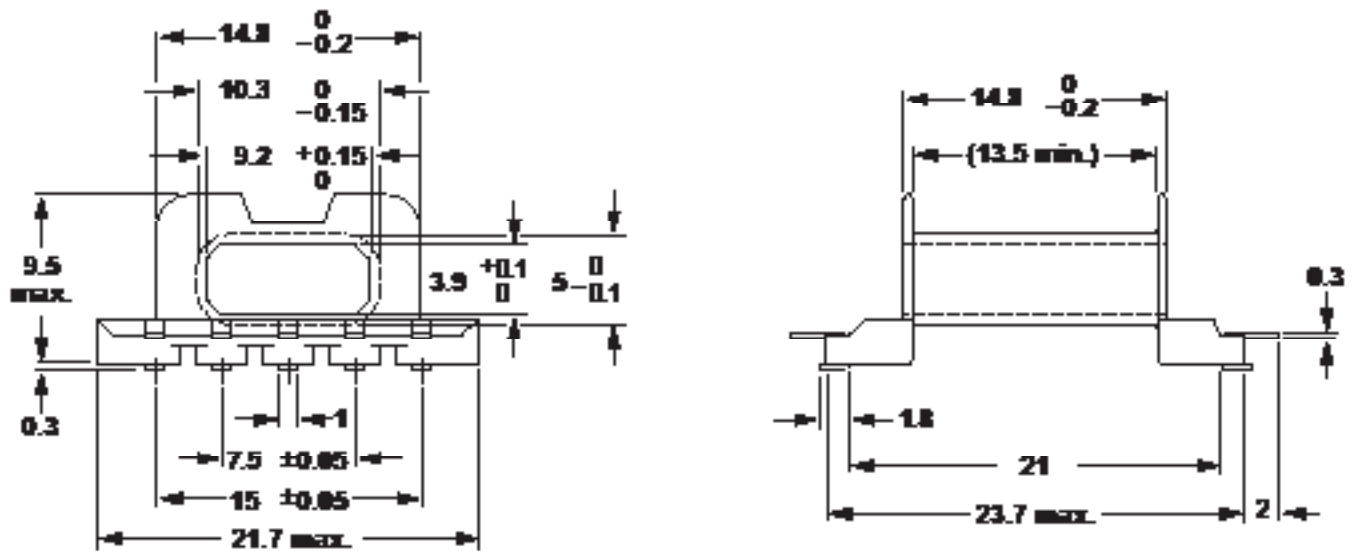
$$\text{Number of available layers} = \frac{\text{Buildup (mm)}}{\text{Diameter of wire (mm)}} \quad (24)$$

The total bobbin turns are calculated by multiplying turns per layer by the number of layers, as in Equation 25.

$$\begin{aligned} \text{Total bobbin turns} &= \quad (25) \\ \text{Turns per layer} \times \text{Number of available layers} \end{aligned}$$

The winding fill factor for a given bobbin is then given in Equation 26. A basic rule of thumb for the winding factor of a given bobbin is in the range of 0.3 to 0.7, depending on isolation requirements. If you need high-voltage isolation, you will need a lower fill factor to fit the extra tape needed between the windings.

$$\text{Winding fill factor} = \frac{\text{Turns needed}}{\text{Total bobbin turns available}} \quad (26)$$



Winding data and area product for EFD20/10/7 coil former (SMD) with 10-solder pads

Number of Sections	Winding Area (mm ²)	Minimum Winding Width (mm)	Average Length of Turn (mm)	Area Product Ae x Aw (mm ²)	Type Number
1	27.7	13.5	34.1	859	CPHS-EFD20-1S-10P

Figure 9 – Manufacturer bobbin data (courtesy of Ferroxcube).

XIV. COUPLED INDUCTOR LAYERS

Transformer layout is very important because it affects primary-to-secondary coupling and leakage inductance. In most applications – especially for the example in this paper – coupling needs to be maximized, which in turn will minimize leakage inductance.

The best coupling is achieved with a one-to-one layout, meaning that the primary and

secondary are wound together as a twisted-pair wire in one layer on a bobbin. This is not usually possible, however, because of the turns ratio needed and the bobbin window size, so interleaving the winding is the next best way to go.

Figure 10 is an example of interleaving a secondary winding between a split primary winding. This approach not only yields good coupling and low leakage, but also lowers proximity effects.

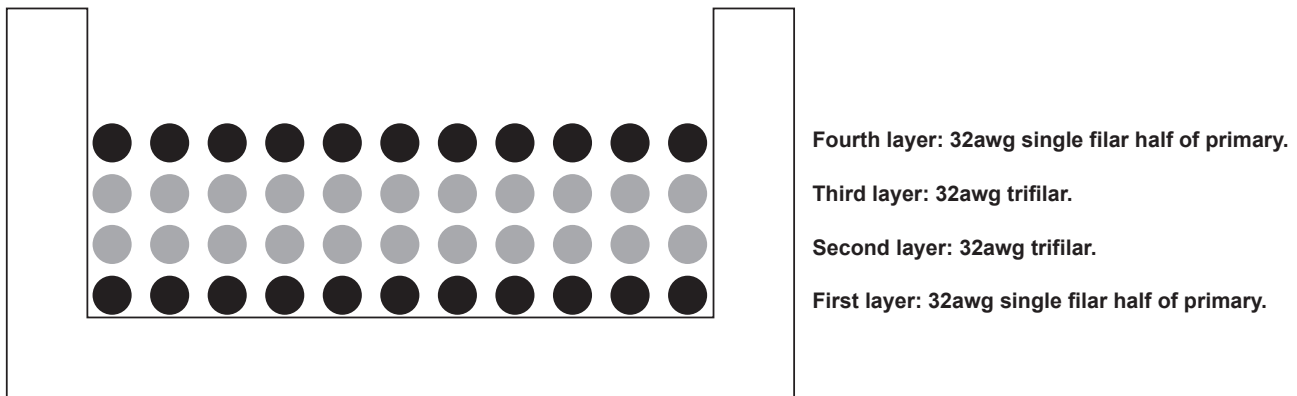


Figure 10 – Interleaved winding layout.

XV. DESIGN FLOW SUMMARY

Here are the steps, along with the design flow diagram (Figure 11) for the coupled-inductor design example:

1. Complete specification.
2. Pick a core (see section 4 of “Power Transformer Design” by Lloyd H. Dixon).
3. Calculate inductance and turns based on Al .
4. Calculate copper loss.
5. Calculate flux density and core loss.
6. Calculate temperature rise (see Robert Kollman’s “Constructing Your Power Supply – Layout Considerations” from the 2002 Power Supply Design Seminar).
7. Reiterate as required to get the turns to layer nicely on the core for good coupling and low leakage inductance.

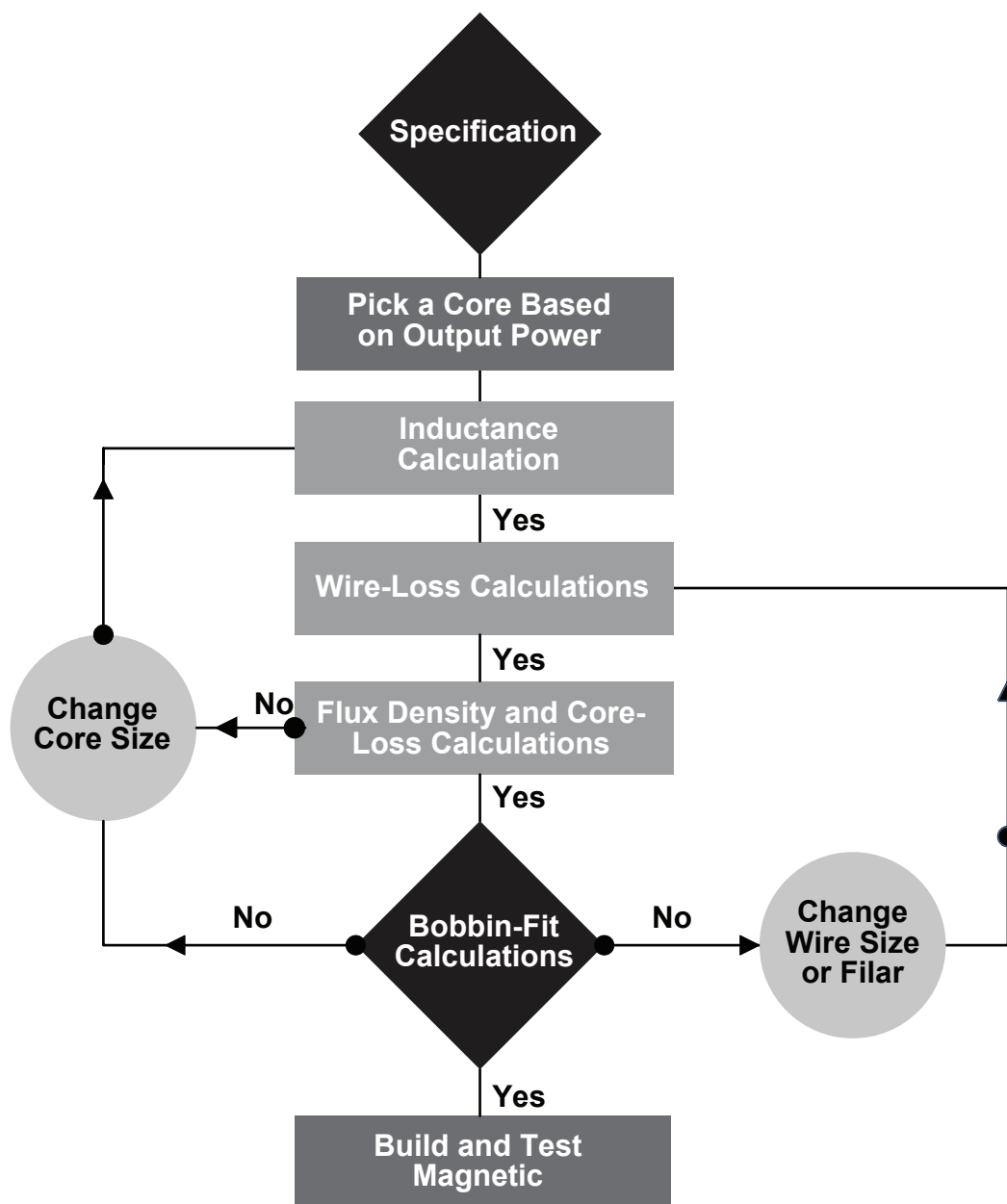


Figure 11 – Design flow diagram.

XVI. FLYBACK-COUPLED INDUCTOR EXAMPLE

A. Step 1: Complete Specification

Before designing a coupled inductor, you will need the information shown in Table 2.

General	
Topology	Quasi-Resonant Flyback
Main Output Power	10 W
Maximum Switching Frequency at Full Load	140 kHz
Input	
Minimum Input Voltage	85 VAC, 76 VDC
Maximum Input Voltage	265 VAC, 375 VDC
Primary Peak Current	1.155 A
Primary RMS Current	0.425 A
Outputs	
Secondary Output Voltage	5 V
Secondary Peak Current	13.861 A
Secondary RMS Current	5.382 A
Bias Voltage	16 V
Bias Current	50 mA
Inductance and Turns Ratio	
Primary Inductance	190.918 μ H
Leakage Inductance	3.818 μ H
Primary to Secondary Turns Ratio	12

Table 2 – Magnetic specification.

B. Step 2: Core Selection

Core selection is based on size, material and gap. These factors work together to determine the flux density, core loss and inductance per turn of the core. Reference [1] has a lot of information about core selection, particularly section 4, “Power Transformer Design.”

Unfortunately, as I said before, there is no replacement for experience when it comes to core selection. Magnetic design is a very iterative process; the more experience you have, the fewer iterations you will have to do.

For this example, I selected a Ferroxcube EFD20 core using 3F3 material. The gap will be set to obtain an $Al = 82 \times 10^{-9}$.

As shown in the data from Ferroxcube in Figure 12, the Al required is not standard; the core gap had to be custom-cut to obtain the necessary Al . The core gap was not determined initially, but only after iteration, by determining the number of turns that would fit in one layer on the bobbin.



GRADE	A_L (nH)	μ_0	TOTAL AIR GAP (μm)	TYPE NUMBER
3C90	83 $\pm 3\%$ ⁽¹⁾	~ 76	~ 960	EFD20/10/7-3C90-E83
	100 $\pm 3\%$	~ 121	~ 510	EFD20/10/7-3C90-A100
	160 $\pm 5\%$	~ 193	~ 280	EFD20/10/7-3C90-A160
	250 $\pm 8\%$	~ 302	~ 160	EFD20/10/7-3C90-A250
	315 $\pm 10\%$	~ 380	~ 120	EFD20/10/7-3C90-A315
	1300 $\pm 25\%$	~ 1570	~ 0	EFD20/10/7-3C90
3C94	83 $\pm 3\%$ ⁽¹⁾	~ 76	~ 960	EFD20/10/7-3C94-E83
	100 $\pm 3\%$	~ 121	~ 510	EFD20/10/7-3C94-A100
	160 $\pm 5\%$	~ 193	~ 280	EFD20/10/7-3C94-A160
	250 $\pm 8\%$	~ 302	~ 160	EFD20/10/7-3C94-A250
	315 $\pm 10\%$	~ 380	~ 120	EFD20/10/7-3C94-A315
	1300 $\pm 25\%$	~ 1570	~ 0	EFD20/10/7-3C94
3C95 	1540 $\pm 25\%$	~ 1865	~ 0	EFD20/10/7-3C95
3C96 	1200 $\pm 25\%$	~ 1450	~ 0	EFD20/10/7-3C96
3F3	83 $\pm 3\%$ ⁽¹⁾	~ 76	~ 960	EFD20/10/7-3F3-E83
	100 $\pm 3\%$	~ 121	~ 510	EFD20/10/7-3F3-A100
	160 $\pm 5\%$	~ 193	~ 280	EFD20/10/7-3F3-A160
	250 $\pm 8\%$	~ 302	~ 160	EFD20/10/7-3F3-A250
	315 $\pm 10\%$	~ 380	~ 120	EFD20/10/7-3F3-A315
	1200 $\pm 25\%$	~ 1450	~ 0	EFD20/10/7-3F3

Figure 12 – Manufacturer core/material data (courtesy of Ferroxcube).

C. Step 3: Calculate Inductance and Turns

Based on Al .

Given that $L_p = 190 \mu\text{H}$, and from iteration, set $Al = 82 \times 10^{-9}$ and solve for N_p :

$$N_p = \left(\frac{L_p}{Al} \right)^{1/2} = 48 \text{ primary turns}$$

Solve for secondary turns:

Turns ratio = 12-1

$$N_s = \frac{N_p}{\text{Turns ratio}} = 4$$

D. Step 4: Calculate Copper Loss

Determine the primary wire area where $J = 400 \text{ A/cm}^2$:

$$\text{Primary wire area} = \frac{I_{\text{primary_rms}}}{J}$$

$$= 1.0625 \times 10^{-3} \text{ cm}^2$$

Determine the skin depth:

$$f = 140 \text{ kHz}$$

$$\text{Skin depth at } 100^\circ\text{C} = \frac{7.6}{\sqrt{f}} = 0.0203 \text{ cm}$$

You will need calculations for a wire gauge range from 26AWG to 32AWG because you have not yet determined the correct wire gauge.

AWG	Copper Diameter in cm	Copper Radius in cm	Copper Area in cm^2	Ω/cm at 100°C
26	0.04	0.02	0.001287	0.001789
28	0.032	0.016	0.00081	0.002845
30	0.025	0.0125	0.000509	0.004523
32	0.02	0.01	0.00032	0.007192

Table 3 – Copper wire data. This information is also in reference [3].

Determine the annular inner ring area in centimeters squared at 100°C :

Keep in mind that when the skin depth is greater than the radius of the wire, you must set the ring area equal to the area of the wire.

$$\text{Area ring 26AWG } 100^{\circ}\text{C} = \text{area 26AWG cm}^2 = 1.287 \times 10^{-3}$$

$$\text{Area ring 28AWG } 100^{\circ}\text{C} = \text{area 28AWG cm}^2 = 8.1 \times 10^{-4}\text{cm}^2$$

$$\text{Area ring 30AWG } 100^{\circ}\text{C} = \text{area 30AWG cm}^2 = 5.09 \times 10^{-4}\text{cm}^2$$

$$\text{Area ring 32AWG } 100^{\circ}\text{C} = \text{area 32AWG cm}^2 = 3.2 \times 10^{-4}\text{cm}^2$$

Determine the ratio of R_{dc} to R_{ac} for a range of wire gauges from 26AWG to 32AWG:

$$R_{\text{skin 26AWG } 100^{\circ}\text{C}} = \frac{\text{Area 26AWG cm}^2}{\text{Area ring 26AWG } 100^{\circ}\text{C cm}^2} = 1$$

$$R_{\text{skin 28AWG } 100^{\circ}\text{C}} = \frac{\text{Area 28AWG cm}^2}{\text{Area ring 28AWG } 100^{\circ}\text{C}} = 1$$

$$R_{\text{skin 30AWG } 100^{\circ}\text{C}} = \frac{\text{Area 30AWG cm}^2}{\text{Area ring 30AWG } 100^{\circ}\text{C}} = 1$$

$$R_{\text{skin 32AWG } 100^{\circ}\text{C}} = \frac{\text{Area 32AWG cm}^2}{\text{Area ring 32AWG } 100^{\circ}\text{C}} = 1$$

Determine wire resistance in ohms per centimeter due to skin effects:

$$R_{\text{copper 26AWG } 100^{\circ}\text{C}} = \text{wire 26AWG ohms/cm } 100^{\circ}\text{C} \times R_{\text{skin 26AWG } 100^{\circ}\text{C}} = 1.789 \times 10^{-3}\text{ohms/cm}$$

$$R_{\text{copper 28AWG } 100^{\circ}\text{C}} = \text{wire 28AWG ohms/cm } 100^{\circ}\text{C} \times R_{\text{skin 28AWG } 100^{\circ}\text{C}} = 2.845 \times 10^{-3}\text{ohms/cm}$$

$$R_{\text{copper 30AWG } 100^{\circ}\text{C}} = \text{wire 30AWG ohms/cm } 100^{\circ}\text{C} \times R_{\text{skin 30AWG } 100^{\circ}\text{C}} = 4.523 \times 10^{-3}\text{ohms/cm}$$

Determine the number of primary wires in parallel required:

$$\text{Numbers of wires primary 26AWG} = \frac{\text{Area required primary cm}^2}{\frac{\text{Area 26AWG cm}}{R_{\text{skin 26AWG } 100^{\circ}\text{C}}}} = 0.826$$

$$\text{Numbers of wires primary 28AWG} = \frac{\text{Area required primary cm}^2}{\frac{\text{Area 28AWG cm}}{R_{\text{skin 28AWG } 100^{\circ}\text{C}}}} = 1.312$$

$$\text{Numbers of wires primary 30AWG} = \frac{\text{Area required primary cm}^2}{\frac{\text{Area 30AWG cm}}{R_{\text{skin 30AWG } 100^{\circ}\text{C}}}} = 2.087$$

Determine primary copper loss:

Number of primary wires = 1

The length per turn in mm from Figure 9 is 34.1 mm:

$$\text{Length per turn cm} = \frac{34.1 \text{ mm}}{10} = 3.41 \text{ cm}$$

$$R_{\text{copper primary resistance 26AWG}} = R_{\text{copper 26AWG } 100^{\circ}\text{C}} \times N_p \times \frac{\text{Length per turn cm}}{\text{Number of primary wires}} = 0.293 \text{ ohms}$$

$$P_{\text{copper primary loss 26AWG}} = I_{\text{primary rms}}^2 \times R_{\text{copper primary 26AWG}} = 0.053 \text{ W}$$

Determine secondary wire AWG and losses:

$$N_s = 4$$

Where:

$$J = 400 \text{ A/cm}^2$$

$$\text{Secondary wire area} = \frac{I_{\text{secondary rms}}}{J} = 14 \times 10^{-3} \text{ cm}^2$$

$$\text{Number of wires secondary 28AWG} = \frac{\text{Secondary wire area cm}^2}{\frac{\text{Area 28AWG cm}}{R_{\text{skin 28AWG } 100^{\circ}\text{C}}}} = 16.667$$

Only five wires of 28AWG in parallel will fit on the chosen bobbin. Through iteration and winding factor calculations, I determined that the number of wires in the secondary = 5.

$$R_{\text{copper secondary 28AWG}} = R_{\text{copper 28AWG } 100^{\circ}\text{C}} \times N_s \times \frac{\text{Length per turn cm}}{\text{Number of wires secondary}} = 7.761 \times 10^{-3} \text{ ohms}$$

$$P_{\text{copper secondary loss 28AWG}} = I_{\text{secondary rms}}^2 \times R_{\text{copper secondary 28AWG}} = .226 \text{ W}$$

Determine bias wire AWG and losses:

Only one wire of 32AWG will fit on the chosen bobbin.

Through iteration and winding factor calculations, I determined that the number of bias wires = 1.

$$\text{Number of turns bias winding (Nsb)} = N_s \times \frac{V_{\text{bias}}}{V_{\text{out}}} \approx 13$$

$$R_{\text{copper bias 32AWG}} = R_{\text{copper 32AWG } 100^{\circ}\text{C}} \times N_{\text{sb}} \times \frac{\text{Length per turn cm}}{\text{Number of bias wires}} = 225 \times 10^{-3} \text{ ohms}$$

$$P_{\text{copper bias loss 32AWG}} = I_{\text{bias rms}}^2 \times R_{\text{copper bias 32AWG}} = 4.197 \times 10^{-4} \text{ W}$$

E. Step 5: Calculate Flux Density and Core Loss

Figure 13 is data from magnetic manufacturer Ferroxcube's data book.

CORES

Effective core parameters

SYMBOL	PARAMETER	VALUE	UNIT
$\Sigma(l/A)$	core factor (C1)	1.52	mm ⁻¹
V_e	effective volume	1460	mm ³
l_e	effective length	47.0	mm
A_e	effective area	31.0	mm ²
A_{min}	minimum area	29	mm ²
m	mass of core half	~3.5	g

Figure 13 – Effective core parameters (courtesy of Ferroxcube).

$$\beta_{ac} = \frac{V_{in\ min} \times T_{on\ max}}{A_e \times N_p} \times 10^8 =$$

$$1.4817 \times 10^3 \text{ Gauss} = 148.17 \text{ mT}$$

$$\beta_{max} = \frac{L_p \times I_{pri_p}}{A_e \times N_p} \times 10^8 =$$

$$1.4817 \times 10^3 \text{ Gauss} = 148.17 \text{ mT}$$

Where:

$$A_e = .31 \text{ cm}^2$$

$$N_p = 48$$

$$I_{pri_p} = 1.155 \text{ A}$$

$$T_{on\ max} = 2.9 \times 10^{-6}$$

For a flyback converter running in discontinuous current mode, the AC flux density, β_{ac} , is equal to the maximum flux density, β_{max} . Figure 14 shows that the saturation point for the given core is about 250 mT. So this design is about 60 percent of maximum flux.

Determine core loss:

$$B_{unipolar} = \frac{\beta_{ac}}{2} = 74.085 \text{ Gauss}$$

$$B_{unipolar\ mT} = \frac{B_{unipolar\ Gauss}}{10} = 74.085 \text{ mT}$$

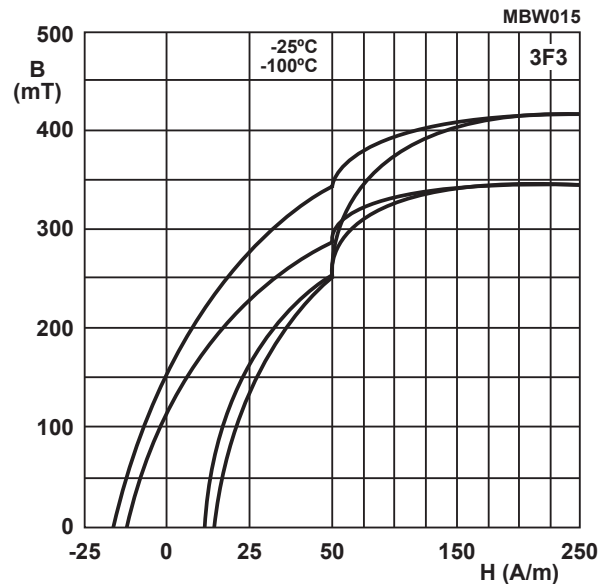


Figure 14 – BH curves (courtesy of Ferroxcube).

From Figure 15, core power dissipation is obtained in kW/m³:

Where:

$$P_{core} = 60 \text{ kW/m}^3$$

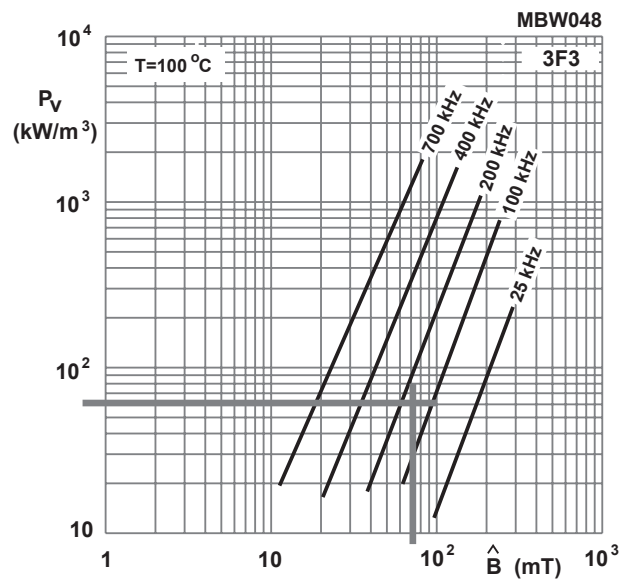


Figure 15 – Specific power loss as a function of peak flux density with frequency as a parameter (courtesy of Ferroxcube).

V_e is given in mm^3 , so you must convert V_e to m^3 :

$$V_e \text{ per m}^3 = \frac{V_e}{1,000^3} = 1.46 \times 10^{-6} \text{ m}^3$$

$$P_{\text{core loss}} = P_{\text{core}} \times V_e \text{ per m}^3 = 88 \text{ mW}$$

Total winding and core power dissipation:

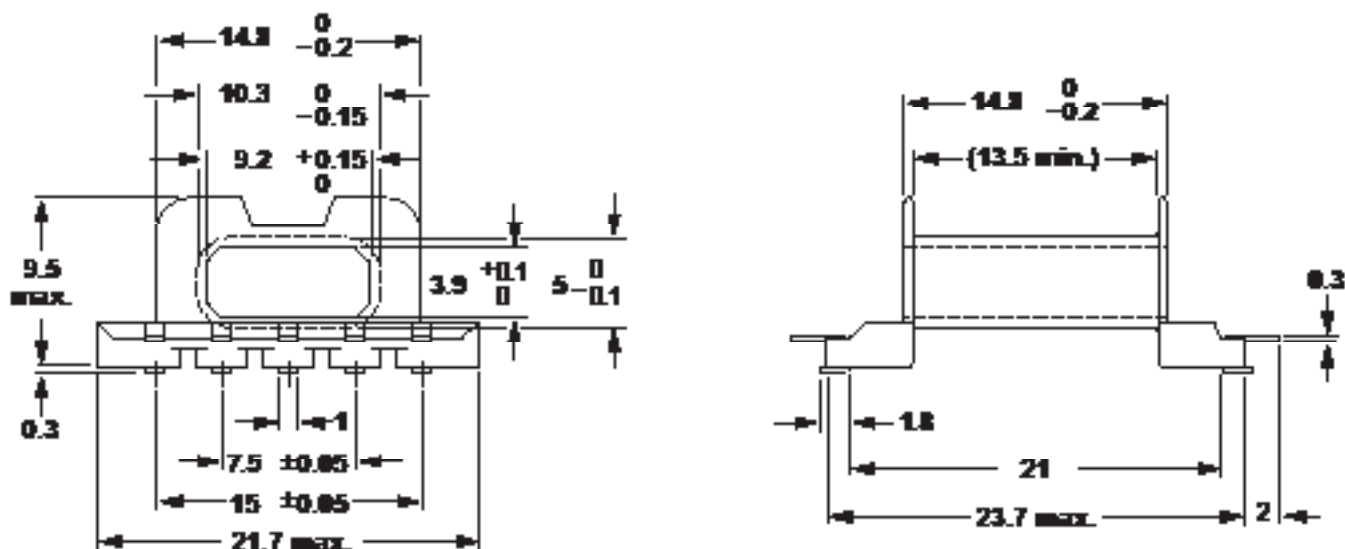
$$\begin{aligned} &\text{Total magnetic power dissipation} \\ &= P_{\text{core loss}} + \text{primary winding loss} \\ &+ \text{secondary winding loss} \\ &+ \text{bias winding loss} = 367 \text{ mW} \end{aligned}$$

F. Step 6: Bobbin Fit Factor

Use Table 4, along with the manufacturer data shown in Figure 16, to calculate the bobbin-fit factor.

AWG	Copper Diameter in cm with Insulation	Copper Area in cm^2 with Insulation
26	0.046	0.001671
28	0.037	0.001083
30	0.03	0.000704
32	0.024	0.000459

Table 4 – Copper wire dimensions.



Winding data and area product for EFD20/10/7 coil former (SMD) with 10-solder pads

Number of Sections	Winding Area (mm^2)	Minimum Winding Width (mm)	Average Length of Turn (mm)	Area Product $A_e \times A_w$ (mm^2)	Type Number
1	27.7	13.5	34.1	859	CPHS-EFD20-1S-10P

Figure 16 – Winding data (courtesy of Ferroxcube).

Bobbin width cm = 1.35 cm

$$\text{Turns per layer 26AWG} = \frac{\text{Bobbin width cm}}{\text{Diameter 26AWG with isolation}} - 2 \approx 27$$

Turns per layer 28AWG = 24 Turns per layer 30 AWG = 42

$$\text{Winding area in cm}^2 = .277 \text{ cm}^2$$

$$\text{Buildup in cm} = \frac{\text{Winding area cm}^2}{\text{Bobbin width cm}} = .205 \text{ cm}$$

$$\text{Layers} = \frac{\text{Buildup in cm}}{\text{Diameter 26AWG with isolation}} \approx 4$$

$$\text{Total bobbin turns 26AWG} = \text{turns per layer} \times \text{layers} = 108$$

Total turns needed

$$= N_p \times \text{number of primary wires} + N_s \times \text{number of secondary wires} \\ + N_{sb} \times \text{number of bias wires} = 81$$

$$\text{Winding factor} = \frac{\text{Total turns needed}}{\text{Total bobbin turns 26AWG}} = .75$$

This is a worst-case winding factor, due to the fact that the secondary windings are made up of 28AWG and 32AWG. On the secondary layer there are 20 turns of 28AWG, which fills 59 percent of the layer, and 13 turns of 32AWG, which fills 24 percent of the layer, for a total fill factor of 83 percent. This should leave enough room for tape.

Figure 17 shows the layout for the coupled inductor. This is an interleaved winding in that the primary is split in to two layers, with the secondary

in the middle. This gives very good coupling and low leakage inductance. The secondary layer has both the secondary winding and the bias winding. The black lines show how the tape should run.

- First layer: 24 turns of 26 AWG single filer, half of primary.
- Second layer: Four turns of 28 AWG five filer secondary plus 13 turns of 32AWG bias winding.
- Third layer: 24 turns of 26 AWG single filer, half of primary.

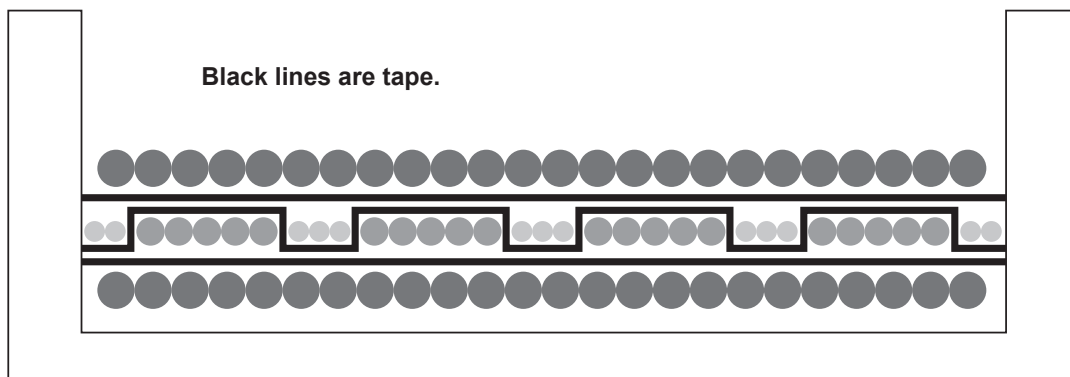


Figure 17 – Coupled inductor layout.

XVII. REFERENCES

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