Design of Flyback Transformers and Filter Inductors

by Lloyd H. Dixon, Jr.

Topic 2

FILTER INDUCTOR AND FLYBACK TRANSFORMER DESIGN FOR SWITCHING POWER SUPPLIES

Lloyd H. Dixon, Jr

This design procedure applies to magnetic devices used primarily to store energy. This includes inductors used for filtering in Buck regulators and for energy storage in Boost circuits, and "flyback transformers" (actually inductors with multiple windings) which provide energy storage, coupling and isolation in Flyback regulators. The design of true transformers used for coupling and isolation in circuits of the Buck and Boost families (in which energy storage is undesired) is covered in Section M5 of this manual.

Symbols, definitions, basic magnetic design equations and various core and wire data used in this section are defined in Reference Sections M1, M2, and M3, and in Appendix A at the end of this section. The specific equations used in this design procedure are derived in Appendix B. The Standard International system of units (rationalized MKS) is used in developing the equations, but dimensions have been converted from meters to centimeters.

All circuit values such as inductance, peak and rms currents and turns ratios must be defined before beginning the magnetics design procedure.

A practical design example of a flyback transformer design using this procedure is in the paper: "150 Watt Flyback Regulator".

Step 1. Select the Core Material and Configuration

Ferrite is the most widely used core material for commercial applications (see Section M3). Molybdenum-permalloy powder toroidal cores have higher losses, but they are often used at switching frequencies below 100 kHz when the flux swing is small -- in filter inductors and flyback transformers operated in the continuous mode. Powdered iron cores are sometimes used, but they are generally either too low in permeability or too lossy for practical use in switching power supply applications above 20 kHz.

The basic magnetic materials above all have very high permeabilities (μ_r = 3000 - 100,000) and cannot therefore store much energy. This is good for a true transformer, but not for an inductor. The large amount of energy that must be stored in a filter inductor or flyback transformer is in fact stored in an air gap (or other non-magnetic material with $\mu_r = 1$) in series with the high permeability core material. In moly-permalloy and powdered iron cores the energy storage gap is actually in the non-magnetic binder holding the magnetic particles together. This distributed gap cannot be measured or specified directly, so the equivalent permeability of the overall composite core is specified instead.

Step 2. Determine the Peak Flux Density

In the following procedure, inductance and current values are referred to the primary. (The single winding of a simple inductor will also be called the primary.) The inductance, L, required and the peak short-circuit

UNITRODE CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

inductor current, I_{pk}, are dictated by the circuit application. I_{pk} is set by the current limiting circuit. Together, these define the absolute maximum inductor energy, $(LI_{pk}^2)/2$, that the inductor must be designed to store (in the gap) without saturating the core and with acceptable core losses and copper losses.

The maximum peak flux density, B_{max}, that will occur at I_{pk} must be defined. The inductor should be operated at B_{max} as large as possible to achieve the smallest possible gap capable of storing the required energy. This minimizes the winding turns, eddy current losses, and inductor size and cost.

In practice, B_{max} is limited either by core saturation, B_{sat}, or by core losses. Core losses in ferrite are proportional to frequency and to the approximate 2.4th power of the peak-to-peak flux density swing, ΔB, during each switching cycle. In an inductor designed to operate in the continuous current mode (such as a buck regulator filter inductor or a continuous mode flyback transformer), core losses are usually negligible at frequencies below 500 kHz because ΔB_m is a small fraction of the DC flux level. In these cases, B_{max} can be almost equal to B_{sat}, with a small safety margin. B_{sat} for most power ferrites such as 3C8 material is above 0.3 Tesla (3000 Gauss), and B_{max} of 0.28 - 0.3 Tesla may be tentatively chosen.

In an inductor designed to operate in the discontinuous mode, flux density swings all the way from zero to B_{max} (flux remnance is negligible because of the gap). Thus the maximum flux density swing, ΔB_m, equals B_{max}. In the discontinuous mode, especially at high frequencies, ΔB_m (and B_{max}) will usually be limited by core losses so that B_{max} will be much less than B_{sat}.

Step 3. Determine Core Size

The core used must be able to store the required peak energy in a small gap without saturating and with acceptable core losses. It must contain the required turns with acceptable winding losses. Core selection can be made through an iterative process involving trial solutions, but Equations 1A and 1B provide an approximation of the core area product, AP, required for the application. (AP = window area A_W, times magnetic cross section A_e). Select the smallest core available from catalog data whose area product exceeds the calculated value.

Equation 1A applies when B_{max} is limited by saturation and Equation 1B when limited by core losses. It may be necessary to try both equations, using the largest resulting AP value. First, the saturation limited case. (Refer to Appendix A for symbol definitions):

$$AP = A_W A_e = \left(\frac{L\, I_{pk}\, I_{FL} \cdot 10^4}{450\, K\, B_{max}}\right)^{1.143} \quad \text{cm}^4$$

With L in Henries, B in Tesla, K = see Table I

Equation 1A is based on copper losses at current density J_{max} resulting in a hot spot temperature rise (at the middle of the center-post) of 30°C. J_{max} is a function of core size:

$$J_{30} = 450\, AP^{-.125} \quad \text{A/cm}^2$$

UNITRODE CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

For the core loss limited case, Equation 1B is also based on a hot spot temperature rise of 30°C, but only half due to copper losses and half core losses.

$$\text{(1B)} \qquad AP = A_W A_e = \left[\frac{L\,\Delta I_m\, I_{FL} \cdot 10^4}{130\,K}\right]^{1.34} \cdot (k_H f + k_E f^2)^{.559} \qquad \text{cm}^4$$

For most power ferrites, hysteresis coefficient $k_H = 4 \cdot 10^{-5}$, eddy current coefficient, $k_E = 4 \cdot 10^{-10}$. Equation 1B is based on operating at a current density, J_{max}, contributing 15°C to the hot spot temperature rise:

$$\text{(2B)} \qquad J_{15} = 318\, AP^{-.125} \qquad \text{A/cm}^2$$

Multiple windings, if any, should be proportioned to operate at the same rms current density to assure uniform power distribution in the windings.

TABLE I -- K Factors

	K_u	K_p	$K=K_u \cdot K_p$
Continuous Buck, Boost Inductor:	0.7	1.0	0.7
Discontinuous Boost Inductor:	0.7	1.0	0.7
Continuous Flyback Transformer:	0.4	0.5	0.2
Discontinuous Flyback Transformer:	0.4	0.5	0.2

Window utilization factor K_u of 0.4 for the flyback transformers in Table I includes insulation to meet VDE line isolation requirements, but does not include a bobbin. K_u should be halved for toriodal cores. The primary area factor K_p of 0.5 is for half of the copper area apportioned to the primary, half to the secondary.

4. Define N

The minimum number of turns is next calculated:

$$N_{min} = \frac{L\, I_{pk}}{B_{max} A_e} \cdot 10^4 \qquad \text{when } B_{sat} \text{ limited}$$

$$N_{min} = \frac{L\, \Delta I_m}{\Delta B_m A_e} \cdot 10^4 \qquad \text{when core loss limited}$$

The actual number of turns is the next possible integer value greater than N_{min}. In a flyback transformer with multiple windings, the primary turns may be constrained to specific multiples such as 22, 44, 66, 88 etc. because of turns ratio considerations. In this case if N_{min} is 36 turns, the smallest possible N is 44 turns. It may be that these additional turns above the minimum will not fit the core unless the actual core area product is sufficiently greater than the minimum AP calculated in Equation 1. For the same inductance, the larger N also results in a B_{max} or ΔB_m less than the original limit, and the core losses will be less. Using Equation 3B with the larger value of N and the actual ΔI_m of the application, calculate the smaller value of ΔB_m and use this to find the actual core losses, P_C, from the core manufacturers core loss tables.

UNITRODE CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

5. Calculate the Gap

The gap length is calculated using the classic inductance formula:

(4A)
$$\ell_g = \frac{\mu_o \mu_r N^2 A_e}{L} \cdot 10^{-2} \quad \text{cm}$$

With Ferrite E-E or pot cores, the gap should be in the center-post only, which requires grinding it to size if not available as a standard part. The grinding operation may be avoided by shimming the core halves apart by approximately half the calculated gap length. This puts half the gap in the center-post, with the other half in the outer legs of the core, assuming the cross section area of the combined outer legs equals the centerpost area. The shimming technique results in considerable external magnetic field -- a possible source of EMI. The effective gap is difficult to calculate and it must be adjusted empirically.

In toroidal cores, the gap is distributed between magnetic particles around the entire core, and is inaccessible. Instead of gap length, the core manufacturer specifies the *equivalent* relative permeability as though the core were made entirely of a homogeneous magnetic material. ℓ_e is the effective magnetic path length around the entire core:

$$\text{Max } \mu_r = \frac{L\ \ell_e}{\mu_o N^2 A_e} \cdot 10^{-2}$$

3. Design the Windings

Calculate the maximum total power dissipation, P_{max}, based on the maximum hot spot temperature rise, ΔT, and core thermal resistance, R_T. Subtract the previously calculated core losses, P_C, to determine the maximum winding losses, P_{Cu}:

(5)
$$P_{Cu} = \Delta T / R_T - P_C \quad \text{W}$$

If thermal resistance of the core used is not known, calculate it from the approximation:

$$R_T = 23\ AP^{-.37} \quad {}^{\circ}\text{C/W}$$

Primary winding loss, P_p, obviously equals P_{Cu} in single winding inductors, *but* P_p equals $P_W/2$ with multiple windings. Calculate the maximum primary resistance, using the maximum rms primary current:

$$R_p = P_p / I_{FL}^2 \quad \Omega$$

Divide R_p by the total length of the primary winding to obtain the maximum resistance/cm of the primary conductor:

(7)
$$R_p/cm = R_p / (N\ \ell_t)$$

Enter the wire tables with this R_p/cm value and find the minimum required wire size and its copper area, A_X. Check the total primary conductor area with N wires to make sure it will fit the area available in the core windo

$$(8) \qquad A_P = N A_X \leqslant K_U K_P A_W$$

If A_P is too large, then a larger core must be used and the procedure repeated from Equation 3A or 3B (or a larger temperature rise must be accepted). If A_P is considerably smaller, it may be desirable to use a smaller core. In multiple winding inductors, do *not* use a wire size larger than Equation 7 requires, or leakage inductance and eddy current losses will increase.

The secondary conductor areas are proportioned to the primary conductor area according to the rms currents in each winding, so that the current densities are the same in all windings.

To obtain good coupling between multiple windings, *each winding* must stretch across the entire breadth of the window (the longer dimension), allowing suitable creepage distance at each end. If the turns in any winding, closely wound, do not extend across the entire available winding breadth, they should be spread out. However, this poorly utilizes the window area and results in high eddy current losses if the wire diameter approaches twice the penetration depth. It is much better to replace a single large diameter conductor with several paralleled conductors which can occupy the available area much more compactly and also reduce eddy current losses.

For example, suppose a tightly wound winding of N turns of diameter D and area A occupies only half the available winding breadth. The height of the winding layer equals D. If this winding is spread out, the coupling to other windings will greatly improve, but the height is still D and it occupies twice the volume that it should. If the single wire is replaced by four wires paralleled, each with area $A/4$, diameter $D/2$ and N turns (close wound adjacent to one another as though they were one wire), they will extend exactly across the winding breadth, with a winding height of only $D/2$, and the eddy current losses and leakage inductance will be greatly reduced. The ultimate of this technique is to use thin copper strip for high current windings that have only one or two turns.

The total rms current, I, in any winding usually has a DC component and ar AC component according to the relationship:

$$= I_{DC}^2 + I_{AC}^2$$

The losses in any winding as calculated earlier are caused by the total rms current flowing through the DC resistance of the winding. However, the AC resistance may be much greater than the DC resistance because skin effect and proximity effect cause the AC current component to flow in only a small portion of the total wire area. The ratio R_{AC}/R_{DC} is the resistance factor, F_R. Eddy current losses result from the *rms AC current component only*, I_{AC}, flowing through the higher effective AC resistance of the wire.

In filter inductors used in buck regulators, Eddy current losses are seldom a problem because the AC current component is so small. I_{AC}^2 is typically 1/200 of I_{DC}^2, so F_R would have to be 200 for the eddy current losses to equal the low frequency losses. In continuous mode flyback transformers, the AC component of total inductor current is small and the core losses are therefore small. However, the AC component in each winding is quite large because the current switches back and forth from primary to secondaries, and eddy current losses are usually significant.

The proximity effect is caused by the AC component of the magnetic field that exists between primary and secondary windings. This AC field induces circulating AC currents within each conductor, adding to the DC current in some areas and subtracting in others and greatly increasing the losses. This effect is combated by using paralleled fine wires or thin copper strips which reduce the circulating currents, and by reducing the magnetic field strength. The latter is accomplished by using a wider window to stretch the windings out, by reducing the number of layers in the windings, and by interleaving -- putting half the primary turns inside and half outside the secondaries. The paper "150 Watt Flyback Regulator" gives a practical example of handling these problems.

The thermal resistance with natural convection cooling, R_T, upon which the hot spot temperature rise depends is probably the weakest approximation used in this procedure. R_T is strongly influenced by the shape of the enclosure in which the transformer is mounted, size and location of cooling vents, horizontal vs. vertical mounting surfaces (chimney effect), and obviously by forced air. As a final check, it is a good idea to attach a fine wire thermocouple to the middle of the centerpost and check the temperature rise under conditions approximating the application.

CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

APPENDIX A. SYMBOL DEFINITIONS

International Standard (SI) units are used except dimensions are converted from meters to centimeters. In flyback transformers or multiple winding inductors, symbols refer to primary winding values.

General:

I_{FL}	total rms primary current at full load
I_{pk}	peak short circuit primary current
I_m	maximum continuous peak-to-peak primary current swing
L	primary winding inductance, Henries
P_{MAX}	total power dissipation
R_T	hot spot thermal resistance, natural convection
ΔT	hot spot temperature rise
AP	core area product = $A_W A_e$, cm^4

Winding Parameters:

A_W	total winding window area in core, cm^2
A_{Cu}	total conductor area - all windings
A_p	conductor area of primary winding = $N A_X$
A_X	conductor area of one primary turn
J_{max}	maximum flux density, A/cm^2
K_u	window utilization factor = A_{Cu}/A_W
K_p	primary factor = A_p/A_{Cu}
K	winding factor = $K_u K_p$
ℓ_t	avg. length of 1 turn (MLT), cm
n	turns ratio
N	number of turns
P_{Cu}	winding losses

Core Parameters:

A_p	Conductor area of primary winding, cm^2
A_e	effective center-post area
B_{sat}	saturation flux density, Tesla
B_{max}	maximum peak flux density
ΔB_m	maximum peak-to-peak flux density swing
k_H	core hysteresis loss coefficient
k_E	core eddy current loss coefficient
ℓ_g	gap length, cm
μ_o	permeability of free space = $4\pi \cdot 10^{-7}$ (SI units)
μ_r	relative permeability
P_C	core losses
V_e	core volume

UNITRODE CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

APPENDIX B. DERIVATION OF EQUATIONS

International Standard (SI) units are used in the initial development of these equations, but dimensions are later changed from meters to centimeters. All values are referred to the primary winding.

Circuit energy equals magnetic energy stored in the gap:

(B1) $$\frac{1}{2} L I^2 = \frac{1}{2} B H A_e \ell_g$$

Ampere's Law applied to the nearly linear field within the gap:

$$N I = H \ell_g$$

Substitute $H\ell_g$ into (Bl and simplifying:

$$L I = B A_e N$$

Solve for N:

$$N = \frac{L I}{B A_e} = \frac{L I_{pk}}{B_{max} A_e} \quad \text{when } B_{sat} \text{ limited}$$

$$N = \frac{L \Delta I}{\Delta B A_e} = \frac{L \Delta I_m}{\Delta B_m A_e} \quad \text{when core loss limited}$$

The primary ampere-turns equals the current density times the total primary conductor area:

$$N I = A_p J = J A_w K$$

(B4) $$N = \frac{A_w J K}{I} = \frac{A_w J_{max} K}{I_{FL}}$$

For the saturation limited case, equating N in (B3A) and (B4)

$$\frac{A_w J_{max} K}{I_{FL}} = \frac{L I_{pk}}{B_{max} A_e}$$

Solve for Area Product and convert dimensions (only) to ce imeters:

(B5) $$AP = A_w A_e = \frac{L I_{pk} I_{FL} \cdot 10^4}{J_{max} K B_{max}} \quad cm^4$$

In the case where core operation is saturation limited, core losses are not dominant and the windings are operated at a current density that will produce a 30°C rise with natural convection cooling, from practice:

(B6) $$J_{30} = 450 \, AP^{-.125} \quad A/cm^2$$

Substitute (B6) into (B5) and solve for Area Product:

(B7) $$AP = A_W A_e = \left(\frac{L\, I_{pk}\, I_{FL} \cdot 10^4}{450\, K\, B_{max}}\right)^{1.143} \quad cm^4$$

In the core loss limited case, equate (B3B) and (B4) and convert dimensions to centimeters:

(B8) $$AP = A_W A_e = \frac{L\, \Delta I_m I_{FL} \cdot 10^4}{J_{max} K\, \Delta B_m} \quad cm^4$$

Assume 15°C temperature rise contribution from core losses, 15°C from the windings operating at a current density of:

(B9) $$J_{15} = 318\, AP^{-.125} \quad A/cm^2$$

J_{15} will be substituted for J_{max} in (B8). First, find ΔB_m value that will result in 15°C rise from core losses. Core losses/cm³ can be calculated from the following empirical formula:

(B10) $$P_C/cm^3 = \Delta B_m^{2.4}(k_H f + k_E f^2)$$

Temperature rise depends upon the core losses/cm³ as well as the core volume and thermal resistance:

(B11) $$\Delta T = 15°C = R_T V_e (P_C/cm^3)$$

Thermal resistance and core volume are related empirically to area product:

(B12) $$R_T \cong 23\, AP^{-.37} \quad °C/W$$

(B13) $$V_e \cong 5.7\, AP^{.68} \quad cm^3$$

Substitute (B10), (B12), and (B13) into (B11) and solve for ΔB_m:

(B14) $$\Delta B_m = \frac{0.405 \cdot AP^{-.129}}{(K_H f + K_E f^2)^{.417}}$$

Finally, substitute (B9) and (B14) into (B8) and solve for the core loss limited Area Product requirement;

(B15) $$AP = A_W A_e = \left(\frac{L\, \Delta I_m\, I_{FL} \cdot 10^4}{130\, K}\right)^{1.34} \cdot (k_H f + k_E f^2)^{.559} \quad cm^4$$

UNITRODE CORPORATION • 5 FORBES ROAD • LEXINGTON, MA 02173 • TEL. (617) 861-6540 • TWX (710) 326-6509 • TELEX 95-1064

IMPORTANT NOTICE

Texas Instruments and its subsidiaries (TI) reserve the right to make changes to their products or to discontinue any product or service without notice, and advise customers to obtain the latest version of relevant information to verify, before placing orders, that information being relied on is current and complete. All products are sold subject to the terms and conditions of sale supplied at the time of order acknowledgment, including those pertaining to warranty, patent infringement, and limitation of liability.

TI warrants performance of its products to the specifications applicable at the time of sale in accordance with TI's standard warranty. Testing and other quality control techniques are utilized to the extent TI deems necessary to support this warranty. Specific testing of all parameters of each device is not necessarily performed, except those mandated by government requirements.

Customers are responsible for their applications using TI components.

In order to minimize risks associated with the customer's applications, adequate design and operating safeguards must be provided by the customer to minimize inherent or procedural hazards.

TI assumes no liability for applications assistance or customer product design. TI does not warrant or represent that any license, either express or implied, is granted under any patent right, copyright, mask work right, or other intellectual property right of TI covering or relating to any combination, machine, or process in which such products or services might be or are used. TI's publication of information regarding any third party's products or services does not constitute TI's approval, license, warranty or endorsement thereof.

Reproduction of information in TI data books or data sheets is permissible only if reproduction is without alteration and is accompanied by all associated warranties, conditions, limitations and notices. Representation or reproduction of this information with alteration voids all warranties provided for an associated TI product or service, is an unfair and deceptive business practice, and TI is not responsible nor liable for any such use.

Resale of TI's products or services with *statements different from or beyond the parameters* stated by TI for that product or service voids all express and any implied warranties for the associated TI product or service, is an unfair and deceptive business practice, and TI is not responsible nor liable for any such use.

Also see: Standard Terms and Conditions of Sale for Semiconductor Products. www.ti.com/sc/docs/stdterms.htm

Mailing Address:

Texas Instruments
Post Office Box 655303
Dallas, Texas 75265

Copyright © 2001, Texas Instruments Incorporated