Filter Inductor Design

A variety of factors constrain the design of a magnetic device. The peak flux density must not saturate the core. The peak ac flux density should also be sufficiently small, such that core losses are acceptably low. The wire size should be sufficiently small, to fit the required number of turns in the core window. Subject to this constraint, the wire cross-sectional area should be as large as possible, to minimize the winding dc resistance and copper loss. But if the wire is too thick, then unacceptable copper losses occur due to the proximity effect. An air gap is needed when the device stores significant energy. But an air gap is undesirable in transformer applications. It should be apparent that, for a given magnetic device, some of these constraints are active while others are not significant.

The objective of this handout is to derive a basic procedure for designing a filter inductor. In Section 1, several types of magnetic devices commonly encountered in converters are investigated, and the relevant design constraints are discussed. The design of filter inductors is then covered in Sections 2 and 3. In the filter inductor application, it is necessary to obtain the required inductance, avoid saturation, and obtain an acceptably low dc winding resistance and copper loss. The geometrical constant K_g is a measure of the effective magnetic size of a core, when dc copper loss and winding resistance are the dominant constraints [1,2]. The discussion here closely follows [3]. Design of a filter inductor involves selection of a core having a K_g sufficiently large for the application, then computing the required air gap, turns, and wire size. Design of transformers and ac inductors, where core loss is significant, is covered in a later handout.

1. Several types of magnetic devices, their *B*-*H* loops, and core vs. copper loss

Filter inductor

A filter inductor employed in a CCM buck converter is illustrated in Fig. 1(a). In this application, the value of inductance L is usually chosen such that the inductor current ripple peak magnitude Δi is a small fraction of the full-load inductor current dc component

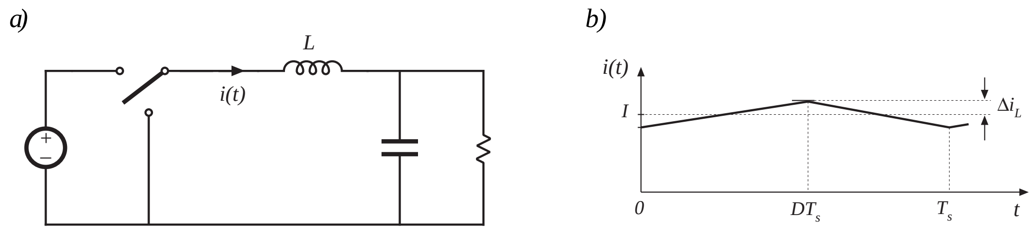


Fig. 1 Filter inductor employed in a CCM buck converter: (a) circuit schematic, (b) inductor current waveform.

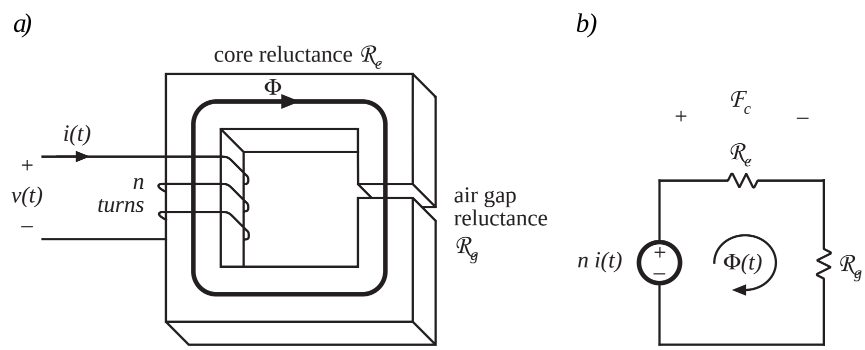


Fig. 2 Filter inductor: (a) structure, (b) magnetic circuit model.

I, as illustrated in Fig. 1(b). As illustrated in Fig. 2, an air gap is employed which is sufficiently large to prevent saturation of the core by the peak current $I + \Delta i$.

The core magnetic field strength $H_c(t)$ is related to the winding current $i(t)$ according to

$$H_c(t) = \frac{n\, i(t)}{l_c} \frac{\mathcal{R}_c}{\mathcal{R}_c + \mathcal{R}_g} \tag{1}$$

where l_c is the magnetic path length of the core. Since $H_c(t)$ is proportional to $i(t)$, $H_c(t)$ can be expressed as a large dc component H_{c0} and a small superimposed ac ripple ΔH_c, where

$$H_{c0} = \frac{nI}{l_c} \frac{\mathcal{R}_c}{\mathcal{R}_c + \mathcal{R}_g}$$
$$\Delta H_c = \frac{n\, \Delta i}{l_c} \frac{\mathcal{R}_c}{\mathcal{R}_c + \mathcal{R}_g} \tag{2}$$

A sketch of $B(t)$ vs. $H_c(t)$ for this application is given in Fig. 3. This device operates with the minor B-H loop illustrated. The size of the minor loop, and hence the core loss, depends on the magnitude of the inductor current ripple Δi. The copper loss depends on the rms inductor current ripple, essentially equal to the dc component I. Typically, the core loss can be ignored, and the design is driven by the copper loss. The maximum flux density is limited by saturation of

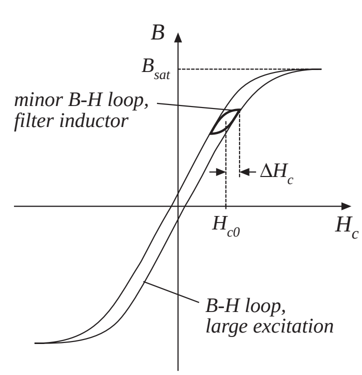


Fig. 3 Filter inductor minor *B*-*H* loop.

the core. Proximity losses are negligible. Although a high-frequency ferrite material can be employed in this application, other materials having higher core losses and greater saturation flux density lead to a physically smaller device.

Ac inductor

An ac inductor employed in a resonant circuit is illustrated in Fig. 4. In this application, the high-frequency current variations are large. In consequence, the *B(t)-H(t)* loop illustrated in Fig. 5 is large. Core loss and proximity loss are usually significant in this application. The maximum flux density is limited by core loss rather than saturation. Both core loss and copper loss must be accounted for in the design of this element. A high-frequency material having low core loss, such as ferrite, is normally employed.

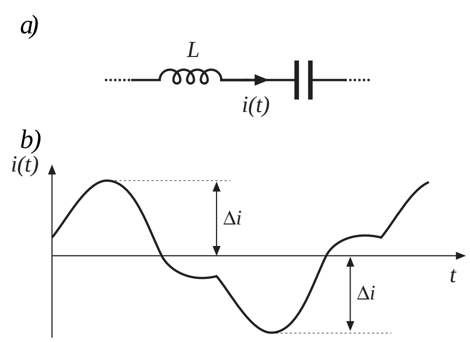


Fig. 4 Ac inductor, resonant converter example: (a) resonant tank circuit, (b) inductor current waveform.

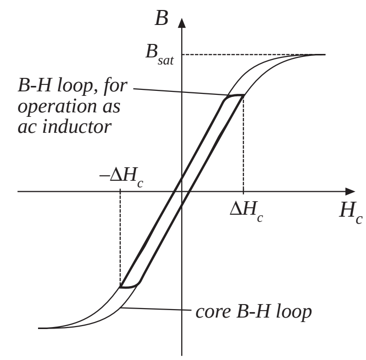


Fig. 5 Operational *B-H* loop of ac inductor.

Transformer

Figure 6 illustrates a conventional transformer employed in a switching converter. Magnetization of the core is modeled by the magnetizing inductance L_{mp}. The magnetizing current $i_{mp}(t)$ is related to the core magnetic field $H(t)$ according to Ampere's law

$$H(t) = \frac{n\, i_{mp}(t)}{l_m} \tag{3}$$

However, $i_{mp}(t)$ is not a direct function of the winding currents $i_1(t)$ or $i_2(t)$. Rather, the

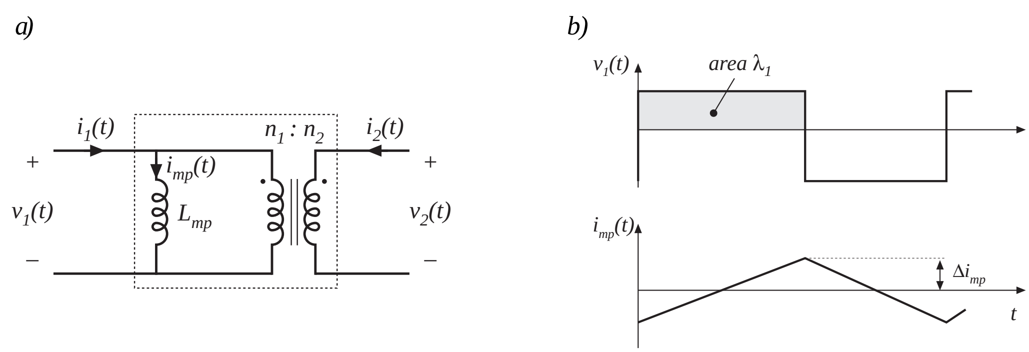


Fig. 6 Conventional transformer: (a) equivalent circuit, (b) typical primary voltage and magnetizing current waveforms.

magnetizing current is dependent on the applied winding voltage waveform $v_1(t)$. Specifically, the maximum ac flux density is directly proportional to the applied volt-seconds λ_1. A typical *B-H* loop for this application is illustrated in Fig. 7.

In the transformer application, core loss and proximity losses are usually significant. Typically the maximum flux density is limited by core loss rather than by saturation. A high-frequency material having low core loss is employed. Both core and copper losses must be accounted for in the design of the transformer.

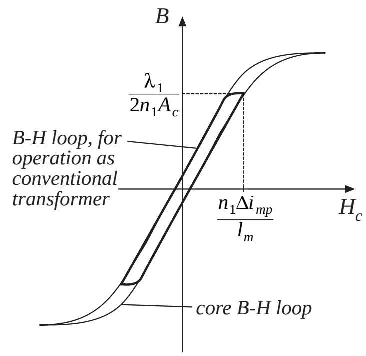


Fig. 7 Operational *B-H* loop of a conventional transformer.

Coupled inductor

A coupled inductor is a filter inductor having multiple windings. Figure 8(a) illustrates coupled inductors in a two-output forward converter. The inductors can be wound on the same core, because the winding voltage waveforms are proportional. The inductors of the SEPIC and Cuk converters, as well as of multiple-output buck-derived converters and some other converters, can be coupled. The inductor current ripples can be controlled by control of the winding leakage inductances [4,5]. Dc currents flow in each winding as illustrated in Fig. 8(b), and the net magnetization of the core is proportional to the sum of the winding ampere-turns:

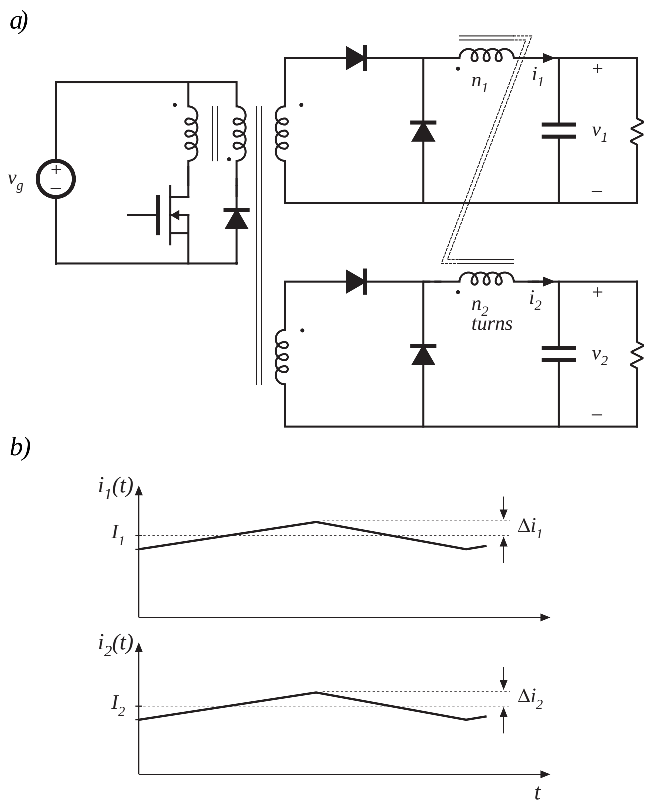


Fig. 8 Coupling the output filter inductors of a two-output forward converter: (a) schematic, (b) typical inductor current waveforms.

$$H_c(t) = \frac{n_1 i_1(t) + n_2 i_2(t)}{l_c} \frac{\mathcal{R}_c}{\mathcal{R}_c + \mathcal{R}_g} \quad (4)$$

As in the case of the single-winding filter inductor, the size of the minor *B-H* loop is proportional to the total current ripple, Fig. 9. Small ripple implies small core loss, as well as small proximity loss. An air gap is employed, and the maximum flux density is limited by saturation.

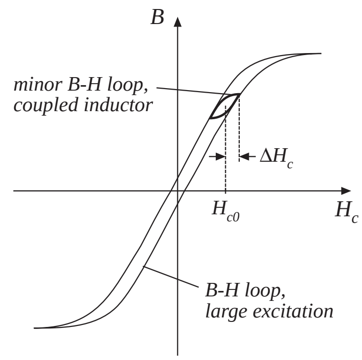


Fig. 9 Coupled inductor minor *B-H* loop.

Flyback transformer

As discussed the Experiment 7 handout, the flyback transformer functions as an inductor with two windings. The primary winding is used during the transistor conduction interval, and the secondary is used during the diode conduction interval. A flyback converter is illustrated in Fig. 10(a), with the flyback transformer modeled as a magnetizing inductance in parallel with an ideal transformer. The magnetizing current $i_{mp}(t)$ is proportional to the core magnetic field strength $H_c(t)$. Typical DCM waveforms are given in Fig. 10(b).

Since the flyback transformer stores energy, an air gap is needed. Core loss depends on the magnitude of the ac component of the magnetizing current. The *B-H* loop for discontinuous conduction mode operation is illustrated in Fig. 11. When the converter is designed to operate in DCM, the core loss is significant. The maximum flux density is then chosen to maintain an acceptably low core loss. For CCM operation, core loss is less significant, and the maximum flux density is limited only by saturation of the core. In either case, winding proximity losses may be of consequence.

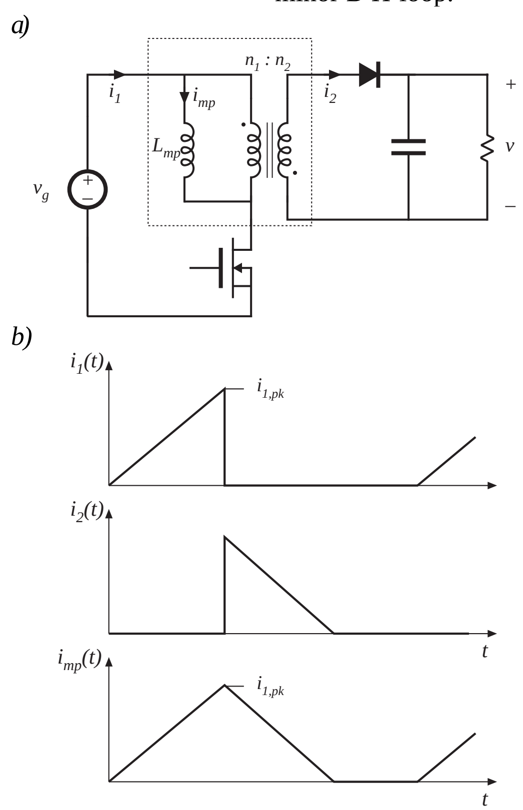


Fig. 10 Flyback transformer: (a) converter schematic, with transformer equivalent circuit, (b) DCM current waveforms.

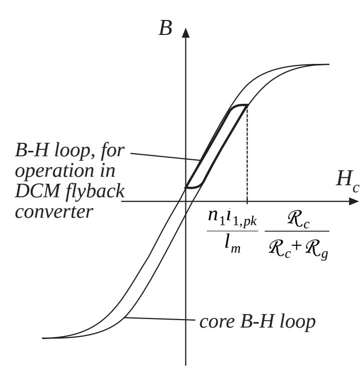


Fig. 11 Operational *B-H* loop of a DCM flyback transformer.

2. Filter inductor design constraints

Let us consider the design of the filter inductor illustrated in Figs. 1 and 2. It is assumed that the core and proximity losses are negligible, so that the inductor losses are dominated by the low-frequency copper losses. The inductor can therefore be modeled by the equivalent circuit of Fig. 12, in which R represents the dc resistance of the winding. It is desired to obtain a given inductance L and given winding resistance R. The inductor should not saturate when a given worst-case peak current I_{max} is applied. Note that specification of R is equivalent to specification of the copper loss P_{cu}, since

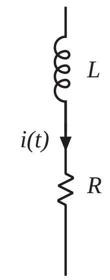


Fig. 12 Filter inductor equivalent circuit.

$$P_{cu} = I_{rms}^{\,2} R \qquad (5)$$

The inductor winding resistance influences both converter efficiency and output voltage. Hence in design of a converter, it is necessary to construct an inductor whose winding resistance is sufficiently small.

It is assumed that the inductor geometry is topologically equivalent to Fig. 13(a). An equivalent magnetic circuit is illustrated in Fig. 13(b). The core reluctance $\mathcal{R}_c$ and air gap reluctance $\mathcal{R}_g$ are

$$\mathcal{R}_c = \frac{l_c}{\mu_c A_c}$$
$$\mathcal{R}_g = \frac{l_g}{\mu_0 A_c} \qquad (6)$$

where l_c is the core magnetic path length, A_c is the core cross-sectional area, μ_c is the core permeability, and l_g is the air gap length. It is assumed that the core and air gap have the same cross-sectional areas. Solution of Fig. 13(b) yields

$$ni = \Phi\left(\mathcal{R}_c + \mathcal{R}_g\right) \qquad (7)$$

Usually, $\mathcal{R}_c << \mathcal{R}_g$, and hence Eq. (7) can be approximated as

$$ni \approx \Phi\,\mathcal{R}_g \qquad (8)$$

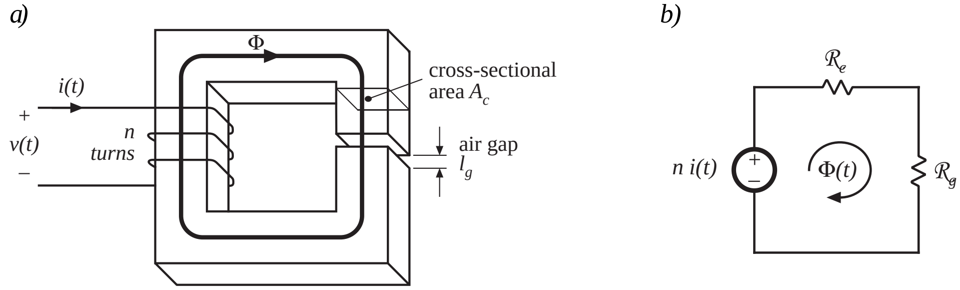


Fig. 13 Filter inductor: (a) assumed geometry, (b) magnetic circuit.

The air gap dominates the inductor properties. Four design constraints now can be identified.

2.1. Maximum flux density

Given a peak winding current I_{max}, it is desired to operate the core flux density at a peak value B_{max}. The value of B_{max} is chosen to be less than the worst-case saturation flux density of the core material.

Substitution of $\Phi = B\,A_c$ into Eq. (8) leads to

$$ni = B\,A_c\,\mathcal{R}_g \tag{9}$$

Upon letting $I = I_{max}$ and $B = B_{max}$, we obtain

$$n\,I_{max} = B_{max}\,A_c\,\mathcal{R}_g = B_{max}\,\frac{l_g}{\mu_0} \tag{10}$$

This is the first design constraint. The turns ratio n and the air gap length l_g are unknowns.

2.2. Inductance

The given inductance value L must be obtained. The inductance is given by

$$L = \frac{n^2}{\mathcal{R}_g} = \frac{\mu_0\,A_c\,n^2}{l_g} \tag{11}$$

This is the second design constraint. The turns ratio n, core area A_c, and gap length l_g are unknown.

2.3. Winding area

As illustrated in Fig. 14, the winding must fit through the window, i.e., the hole in the center of the core. The cross-sectional area of the conductor, or bare area, is A_W. If the winding has n turns, then the area of copper conductor in the window is

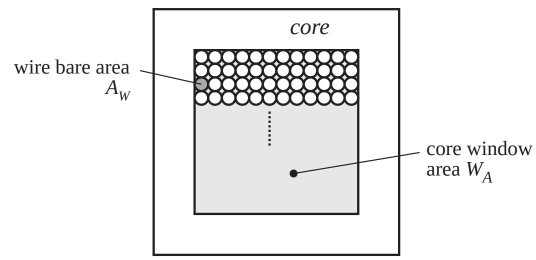


Fig. 14 The winding must fit in the core window area.

$$n\,A_W \tag{12}$$

If the core has window area W_A, then we can express the area available for the winding conductors as

$$W_A\,K_u \tag{13}$$

where K_u is the *window utilization factor*, or *fill factor*. Hence, the third design constraint can be expressed as

$$W_A\,K_u \geq n\,A_W \tag{14}$$

The fill factor K_u is the fraction of the core window area that is filled with copper. K_u must lie between zero and one. As discussed in [1], there are several mechanism which cause K_u to be less than unity. Round wire does not pack perfectly; this reduces K_u by a factor of 0.7 to 0.55, depending on the winding technique. The wire has insulation; the ratio of wire conductor area to total wire area varies from approximately 0.95 to 0.65, depending on the wire size and type of insulation. The bobbin uses some of the window area. Insulation may be required between windings and/or winding layers. Typical values of K_u for cores with winding bobbins are: 0.5 for a simple low-voltage inductor, 0.25-0.3 for an off-line transformer, 0.05-0.2 for a high-voltage transformer supplying several kV, and 0.65 for a low-voltage foil transformer or inductor.

2.4. Winding resistance

The resistance of the winding is

$$R = \rho \frac{l_W}{A_W} \tag{15}$$

where ρ is the resistivity of the conductor material, l_W is the length of the wire, and A_W is the wire bare area. The resistivity of copper at room temperature is $1.724 \cdot 10^{-6}$ Ω-cm. The length of the wire comprising an n-turn winding can be expressed as

$$l_W = n\,(MLT) \tag{16}$$

where *(MLT)* is the mean-length-per-turn of the winding. The mean-length-per-turn is a function of the core geometry. Substitution of Eq. (16) into (15) leads to

$$R = \rho \frac{n\,(MLT)}{A_W} \tag{17}$$

This is the fourth constraint.

3. The core geometrical constant K_g

The four constraints, Eqs. (10), (11), (14), and (17), involve the quantities A_c, W_A, and MLT, which are functions of the core geometry, the quantities I_{max}, B_{max}, μ_0, L, K_u, R, and ρ, which are given specifications or other known quantities, and n, l_g, and A_W, which are unknowns. Elimination of the unknowns n, l_g, and A_W leads to the following equation:

$$\frac{A_c^2\, W_A}{(MLT)} \geq \frac{L^2\, I_{max}^2\, \rho}{B_{max}^2\, R\, K_u} \tag{18}$$

The quantities on the right side of this equation are specifications or other known quantities. The left side of the equation is a function of the core geometry alone. It is necessary to choose a core whose geometry satisfies Eq. (18).

The quantity

$$K_g = \frac{A_c^2 W_A}{(MLT)} \tag{19}$$

is called the core geometrical constant. It is a figure-of-merit which describes the effective electrical size of magnetic cores, in applications where copper loss and maximum flux density are specified. Tables are included in the Appendix which list the values of K_g for several standard families of ferrite cores. K_g has dimensions of length to the fifth power.

Equation (18) reveals how the specifications affect the core size. Increasing the inductance or peak current requires an increase in core size. Increasing the peak flux density allows a decrease in core size, and hence it is advantageous to use a core material which exhibits a high saturation flux density. Allowing a larger winding resistance R, and hence larger copper loss, leads to a smaller core. Of course, the increased copper loss and smaller core size will lead to a higher temperature rise, which may be unacceptable. The fill factor K_u also influences the core size.

Equation (20) reveals how core geometry affects the core capabilities. An inductor capable of meeting increased electrical requirements can be obtained by increasing either the core area A_c, or the window area W_A. Increase of the core area requires additional iron core material. Increase of the window area implies that additional copper winding material is employed. We can trade iron for copper, or vice-versa, by changing the core geometry in a way that maintains the K_g of Eq. (19).

4. A step-by-step procedure

The procedure developed in Section 3 is summarized below. This simple filter inductor design procedure should be regarded as a first-pass approach. Numerous issues have been neglected, including detailed insulation requirements, conductor eddy current losses, temperature rise, roundoff of number of turns, etc.

The following quantities are specified, using the units noted:

Wire resistivity	ρ	(Ω-cm)
Peak winding current	I_{max}	(A)
Inductance	L	(H)
Winding resistance	R	(Ω)
Winding fill factor	K_u	
Core maximum flux density	B_{max}	(T)

The core dimensions are expressed in cm:

core cross-sectional area	A_c	(cm^2)
core window area	W_A	(cm^2)
mean length per turn	MLT	(cm)

The use of centimeters rather than meters requires that appropriate factors be added to the design equations.

Procedure

1. *Determine core size*

$$K_g \ge \frac{L^2 I_{max}^2 \rho}{B_{max}^2 R K_u} 10^8 \qquad (\mathrm{cm}^5) \tag{20}$$

Choose a core which is large enough to satisfy this inequality. Note the values of A_c, W_A, and MLT for this core.

2. *Determine air gap length*

$$l_g = \frac{\mu_0 L I_{max}^2}{B_{max}^2 A_c} 10^4 \qquad (\mathrm{m}) \tag{21}$$

with A_c expressed in cm^2. $\mu_0 = 4\pi \cdot 10^{-7}$ H/m. The air gap length is given in meters. The value expressed in Eq. (13-21) is approximate, and neglects fringing flux and other nonidealities.

Core manufacturers sell gapped cores. Rather than specifying the air gap length, the equivalent quantity A_L is used. A_L is equal to the inductance, in mH, obtained with a winding of 1000 turns. When A_L is specified, it is the core manufacturer's responsibility to obtain the correct gap length. Equation (13-21) can be modified to yield the required A_L, as follows:

$$A_L = \frac{10 B_{max}^2 A_c^2}{L I_{max}^2} \qquad (\mathrm{mH / 1000\ turns}) \tag{22}$$

where A_c is given in cm^2, L is given in Henries, and B_{max} is given in Tesla.

3. *Determine number of turns*

$$n = \frac{L I_{max}}{B_{max} A_c} 10^4 \tag{23}$$

4. *Evaluate wire size*

$$A_W \le \frac{K_u W_A}{n} \tag{24}$$

Select wire with bare copper area less than or equal to this value. An American Wire Gauge table is included in the Appendix.

As a check, the winding resistance can be computed:

$$R = \frac{\rho\, n\, (MLT)}{A_w} \tag{25}$$

5. Summary of key points

1. A variety of magnetic devices are commonly used in switching converters. These devices differ in their core flux density variations, as well as in the magnitudes of the ac winding currents. When the flux density variations are small, core loss can be neglected. Alternatively, a low-frequency material can be used, having higher saturation flux density.
2. The core geometrical constant K_g is a measure of the magnetic size of a core, for applications in which copper loss is dominant. In the K_g design method, flux density and total copper loss are specified.

REFERENCES

[1] C. W. T. McLyman, *Transformer and Inductor Design Handbook,* second edition, New York: Marcel Dekker, 1988.

[2] S. Cuk, "Basics of Switched-Mode Power Conversion: Topologies, Magnetics, and Control," in *Advances in Switched-Mode Power Conversion*, vol. 2, Irvine: Teslaco, pp. 292-305, 1983.

[3] R. W. Erickson, *Fundamentals of Power Electronics*, New York: Chapman and Hall, 1997, Chapter 13.

PROBLEMS

1. A simple buck converter operates with a 50kHz switching frequency and a dc input voltage of V_g = 40V. The output voltage is V = 20V. The load resistance is $R \ge 4\Omega$.

(a) Determine the value of the output filter inductance L such that the peak-to-average inductor current ripple Δi is 10% of the dc component I.

(b) Determine the peak steady-state inductor current I_{max}.

(c) Design an inductor which has the values of L and I_{max} from parts (a) and (b). Use a ferrite E-E core, with B_{max} = 0.25T. Choose a value of winding resistance such that the inductor copper loss is less than or equal to 1W. Assume K_u = 0.5. Specify: core size, gap length, wire size (AWG), and number of turns.

2. A boost converter operates at the following quiescent point: V_g = 28V, V = 48V, P_{load} = 150W, f_s = 100kHz. Design the inductor for this converter. Choose the inductance value such that the peak current ripple is 10% of the dc inductor current. Use a peak flux density of 0.225T, and assume a fill factor of 0.5. Allow copper loss equal to 0.5% of the load power. Use a ferrite PQ core. Specify: core size, air gap length, wire gauge, and number of turns.

3. Coupled inductor design. The two-output forward converter of Fig. 8(a) employs secondary-side inductors which are wound on a common core. An air gap is employed. Winding currents $i_1(t)$ and $i_2(t)$ exhibit coincident peak currents I_{1pk} and I_{2pk}, and rms currents I_{1rms} and I_{2rms}, respectively. The winding resistances are R_1 and R_2.

(a) How should the turns ratio n_2 / n_1 be chosen?

(b) For a given core window area W_A and given I_{1rms} and I_{2rms}, what ratio R_1 / R_2 minimizes the total copper loss? How should the window area be allocated between the windings?

(c) Modify the K_g design procedure, to design this two-winding coupled inductor. Specifications are given regarding the total copper loss P_{cu}, maximum flux density B_{max}, peak and rms winding currents, wire resistivity, fill factor, and converter output voltages. Your procedure should yield the following data: required core geometrical constant K_g, air gap length, primary and secondary turns n_1 and n_2, and primary and secondary wire areas A_{w1} and A_{w2}.

(d) Design a coupled inductor for the following application: V_1 = 5V, V_2 = 15V, I_1 = 20A, I_2 = 4A, D = 0.4. The magnetizing inductance should be equal to 8μH, referred to the 5V winding. You may assume a fill factor K_u of 0.5. Allow a total of 1W of copper loss, and use a peak flux density of B_{max} = 0.2T. Use a ferrite EE core. Specify: core size, air gap length, number of turns and wire gauge for each winding.

Appendix

Magnetics Design Tables

Geometrical data for several standard ferrite core shapes are listed here. The geometrical constant K_g is a measure of core size, useful for designing inductors and transformers that attain a given copper loss. The K_g method for inductor design is described in in this text. K_g is defined as

$$K_g = \frac{A_c^2 W_A}{MLT} \tag{A.1}$$

where A_c is the core cross-sectional area, W_A is the window area, and MLT is the winding mean-length-per-turn. The geometrical constant K_{gfe} is a similar measure of core size, which is useful for designing ac inductors and transformers when the total copper plus core loss is constrained. The K_{gfe} method for magnetics design is described in the handout on transformer design. K_{gfe} is defined as

$$K_{gfe} = \frac{W_A \; A_c^{2(1-1/\beta)}}{MLT \; l_e^{2/\beta}} \; u(\beta) \tag{A.2}$$

where l_e is the core mean magnetic path length, and β is the core loss exponent:

$$P_{fe} = K_{fe} B_{\max}^{\beta} \tag{A.3}$$

For modern ferrite materials, β typically lies in the range 2.6 to 2.8. The quantity $u(\beta)$ is defined as

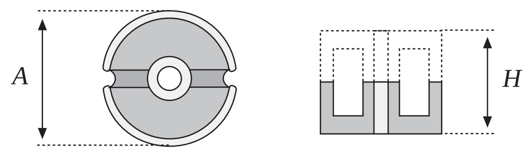
$$u(\beta) = \left[\left(\frac{\beta}{2}\right)^{-\left(\frac{\beta}{\beta+2}\right)} + \left(\frac{\beta}{2}\right)^{\left(\frac{2}{\beta+2}\right)} \right]^{-\left(\frac{\beta+2}{\beta}\right)} \tag{A.4}$$

$u(\beta)$ is equal to 0.305 for $\beta = 2.7$. This quantity varies by roughly 5% over the range $2.6 \le \beta \le 2.8$. Values of K_{gfe} are tabulated for $\beta = 2.7$; variation of K_{gfe} over the range $2.6 \le \beta \le 2.8$ is typically quite small.

Thermal resistances are listed in those cases where published manufacturer's data is available. The thermal resistances listed are the approximate temperature rise from the center leg of the core to ambient, per watt of total power loss. Different temperature rises may be observed under conditions of forced air cooling, unusual power loss distributions, etc. Listed window areas are the winding areas for conventional single-section bobbins.

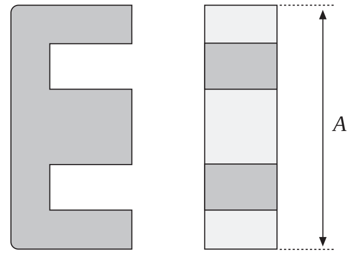
An American Wire Gauge table is included at the end of this appendix.

A.1 Pot core data



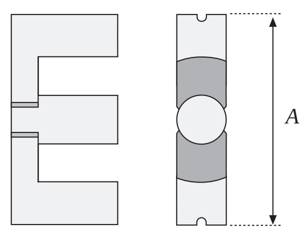
Core type (AH) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (°C/W)	Core weight (g)
704	$0.738\cdot10^{-6}$	$1.61\cdot10^{-6}$	0.070	$0.22\cdot10^{-3}$	1.46	1.0		0.5
905	$0.183\cdot10^{-3}$	$256\cdot10^{-6}$	0.101	0.034	1.90	1.26		1.0
1107	$0.667\cdot10^{-3}$	$554\cdot10^{-6}$	0.167	0.055	2.30	1.55		1.8
1408	$2.107\cdot10^{-3}$	$1.1\cdot10^{-3}$	0.251	0.097	2.90	2.00	100	3.2
1811	$9.45\cdot10^{-3}$	$2.6\cdot10^{-3}$	0.433	0.187	3.71	2.60	60	7.3
2213	$27.1\cdot10^{-3}$	$4.9\cdot10^{-3}$	0.635	0.297	4.42	3.15	38	13
2616	$69.1\cdot10^{-3}$	$8.2\cdot10^{-3}$	0.948	0.406	5.28	3.75	30	20
3019	0.180	$14.2\cdot10^{-3}$	1.38	0.587	6.20	4.50	23	34
3622	0.411	$21.7\cdot10^{-3}$	2.02	0.748	7.42	5.30	19	57
4229	1.15	$41.1\cdot10^{-3}$	2.66	1.40	8.60	6.81	13.5	104

A.2 EE core data



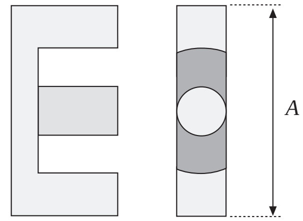
Core type (A) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Core weight (g)
EE12	$0.731\cdot10^{-3}$	$0.458\cdot10^{-3}$	0.14	0.085	2.28	2.7	2.34
EE16	$2.02\cdot10^{-3}$	$0.842\cdot10^{-3}$	0.19	0.190	3.40	3.45	3.29
EE19	$4.07\cdot10^{-3}$	$1.3\cdot10^{-3}$	0.23	0.284	3.69	3.94	4.83
EE22	$8.26\cdot10^{-3}$	$1.8\cdot10^{-3}$	0.41	0.196	3.99	3.96	8.81
EE30	$85.7\cdot10^{-3}$	$6.7\cdot10^{-3}$	1.09	0.476	6.60	5.77	32.4
EE40	0.209	$11.8\cdot10^{-3}$	1.27	1.10	8.50	7.70	50.3
EE50	0.909	$28.4\cdot10^{-3}$	2.26	1.78	10.0	9.58	116
EE60	1.38	$36.4\cdot10^{-3}$	2.47	2.89	12.8	11.0	135
EE70/68/19	5.06	$127\cdot10^{-3}$	3.24	6.75	14.0	9.0	280

A.3 EC core data



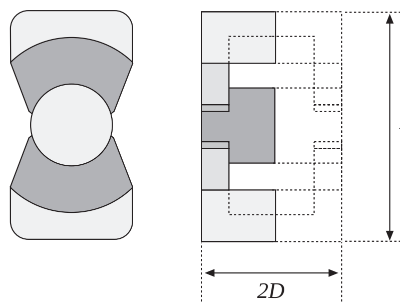
Core type (A) (mm)	Geometrica l constant K_g cm^5	Geometrica l constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (°C/W)	Core weight (g)
EC35	0.131	$9.9\cdot10^{-3}$	0.843	0.975	5.30	7.74	18.5	35.5
EC41	0.374	$19.5\cdot10^{-3}$	1.21	1.35	5.30	8.93	16.5	57.0
EC52	0.914	$31.7\cdot10^{-3}$	1.80	2.12	7.50	10.5	11.0	111
EC70	2.84	$56.2\cdot10^{-3}$	2.79	4.71	12.9	14.4	7.5	256

A.4 ETD core data



Core type (A) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Thermal resistance R_{th} (˚C/W)	Core weight (g)
ETD29	0.0978	$8.5\cdot10^{-3}$	0.76	0.903	5.33	7.20		30
ETD34	0.193	$13.1\cdot10^{-3}$	0.97	1.23	6.00	7.86	19	40
ETD39	0.397	$19.8\cdot10^{-3}$	1.25	1.74	6.86	9.21	15	60
ETD44	0.846	$30.4\cdot10^{-3}$	1.74	2.13	7.62	10.3	12	94
ETD49	1.42	$41.0\cdot10^{-3}$	2.11	2.71	8.51	11.4	11	124

A.5 PQ core data



Core type (A_1/2D) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfe} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_m (cm)	Core weight (g)
PQ 20/16	$22.4\cdot10^{-3}$	$3.7\cdot10^{-3}$	0.62	0.256	4.4	3.74	13
PQ 20/20	$33.6\cdot10^{-3}$	$4.8\cdot10^{-3}$	0.62	0.384	4.4	4.54	15
PQ 26/20	$83.9\cdot10^{-3}$	$7.2\cdot10^{-3}$	1.19	0.333	5.62	4.63	31
PQ 26/25	0.125	$9.4\cdot10^{-3}$	1.18	0.503	5.62	5.55	36
PQ 32/20	0.203	$11.7\cdot10^{-3}$	1.70	0.471	6.71	5.55	42
PQ 32/30	0.384	$18.6\cdot10^{-3}$	1.61	0.995	6.71	7.46	55
PQ 35/35	0.820	$30.4\cdot10^{-3}$	1.96	1.61	7.52	8.79	73
PQ 40/40	1.20	$39.1\cdot10^{-3}$	2.01	2.50	8.39	10.2	95

A.6 American wire gauge data

AWG#	Bare area, 10^{-3} cm^2	Resistance, 10^{-6} Ω/cm	Diameter, cm
0000	1072.3	1.608	1.168
000	850.3	2.027	1.040
00	674.2	2.557	0.927
0	534.8	3.224	0.825
1	424.1	4.065	0.735
2	336.3	5.128	0.654
3	266.7	6.463	0.583
4	211.5	8.153	0.519
5	167.7	10.28	0.462
6	133.0	13.0	0.411
7	105.5	16.3	0.366
8	83.67	20.6	0.326
9	66.32	26.0	0.291
10	52.41	32.9	0.267
11	41.60	41.37	0.238
12	33.08	52.09	0.213
13	26.26	69.64	0.190
14	20.02	82.80	0.171
15	16.51	104.3	0.153
16	13.07	131.8	0.137
17	10.39	165.8	0.122
18	8.228	209.5	0.109
19	6.531	263.9	0.0948
20	5.188	332.3	0.0874
21	4.116	418.9	0.0785
22	3.243	531.4	0.0701
23	2.508	666.0	0.0632
24	2.047	842.1	0.0566
25	1.623	1062.0	0.0505
26	1.280	1345.0	0.0452
27	1.021	1687.6	0.0409
28	0.8046	2142.7	0.0366
29	0.6470	2664.3	0.0330
30	0.5067	3402.2	0.0294
31	0.4013	4294.6	0.0267
32	0.3242	5314.9	0.0241
33	0.2554	6748.6	0.0236
34	0.2011	8572.8	0.0191
35	0.1589	10849	0.0170
36	0.1266	13608	0.0152
37	0.1026	16801	0.0140
38	0.08107	21266	0.0124
39	0.06207	27775	0.0109
40	0.04869	35400	0.0096
41	0.03972	43405	0.00863
42	0.03166	54429	0.00762
43	0.02452	70308	0.00685