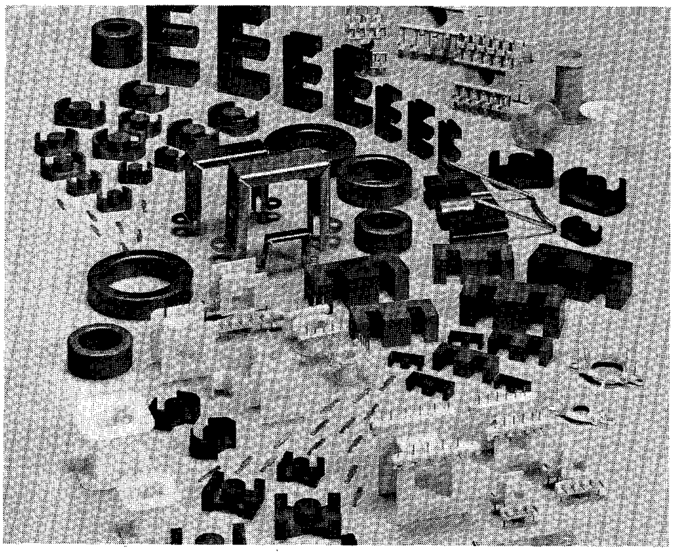
Section 4. Power Design



INTRODUCTION

Ferrite is an ideal core material for transformers, inverters and inductors in the frequency range 20 kHz to 3 MHz, due to the combination of low core cost and low core losses. Tape wound cores do offer higher flux densities and better temperature stability, advantages which may off-set their higher cost.

Ferrite is an excellent material for high frequency (20 kHz to 3 MHz) inverter power supplies. Ferrites may be used in the saturating mode for low power, low frequency operation (<50 watts and 10 kHz). For high power operation a two transformer design, using a tape wound core as the saturating core and a ferrite core as the output transformer, offers maximum performance. The two transformer design offers high efficiency excellent frequency stability, and low switching losses.

Ferrite cores may also be used in fly-back transformer designs, which offer low core cost, low circuit cost and high voltage capability. Powder cores (MPP, High Flux, Kool Mμ®) offer soft saturation, higher B_{max} and better temperature stability and may be the best choice in some flyback applications or inductors.

High frequency power supplies, both inverters and converters, offer lower cost, and lower weight and volume than conventional 60 hertz and 400 hertz power sources.

Many cores in this section are standard types commonly used in the industry. If a suitable size for your application is not listed, Magnetics will be happy to review your needs, and, if necessary, quote tooling where quantities warrant.

Cores are available gapped to avoid saturation under dc bias conditions. J and W materials are available with lapped surfaces.

Bobbins for many cores are available from Magnetics. VDE requirements have been taken into account in bobbin designs for EC, PQ, and metric E Cores. Many bobbins are also available commercially.

CORE SELECTION

Core Materials

F, P, K and R materials, offering the lowest core losses and highest saturation flux density, are most suitable for high power/high temperature operation. P material core losses decrease with temperature up to 70°C; R material losses decrease up to 100°C. K material is recommended for frequencies over 700kHz.

J and W materials offer high impedance for broadband transformers, and are also suitable for low-level power transformers.

**FERRITE
POWER MATERIALS SUMMARY**

		F	P	R	K	J	W +
μi(20)gauss)	25°C	3000	2500	2300	1500	5000	10,000
μp(2000 gauss)	100°C	4600	6500	6500	3500	5500	12,000
Saturation	25°C	4900	5000	5000	4600	4300	4300
Flux Density B_m Gauss	100°C	3700	3900	3700	3900	2500	2500
Core Loss (mw/cm³)	25°C	100	125	140			
(Typical)	60°C	180	80*	100			
@100 kHz, 1000 Gauss	100°C	225	125	70			

*@80°C +@10kHz

CORE GEOMETRIES

Pot Cores

Pot Cores, when assembled, nearly surround the wound bobbin. This aids in shielding the coil from pickup of EMI from outside sources. The pot core dimensions all follow IEC standards so that there is interchangeability between manufacturers. Both plain and printed circuit bobbins are available, as are mounting and assembly hardware. Because of its design, the pot core is a more expensive core than other shapes of a comparable size. Pot cores for high power applications are not readily available.

Double Slab and RM Cores

Slab-sided solid center post cores resemble pot cores, but have a section cut off on either side of the skirt. Large openings allow large size wires to be accommodated and assist in removing heat from the assembly. RM cores are also similar to pot cores, but are designed to minimize board space, providing at least a 40% savings in mounting area. Printed circuit or plain bobbins are available. Simple one piece clamps allow simple assembly. Low profile is possible. The solid center post generates less core loss and this minimizes heat buildup.

E Cores

E cores are less expensive than pot cores, and have the advantages of simple bobbin winding plus easy assembly. Gang winding is possible for the bobbins used with these cores. E cores do not, however, offer self-shielding. Lamination size E shapes are available to fit commercially available bobbins previously designed to fit the strip stampings of standard lamination sizes. Metric and DIN sizes are also available. E cores can be pressed to different thickness, providing a selection of cross-sectional areas. Bobbins for these different cross sectional areas are often available commercially.

E cores can be mounted in different directions, and if desired, provide a low-profile. Printed circuit bobbins are available for low-profile mounting. E cores are popular shapes due to their lower cost, ease of assembly and winding, and the ready availability of a variety of hardware.

EC, ETD and EER Cores

These shapes are a cross between E cores and pot cores. Like E cores, they provide a wide opening on each side. This gives adequate space for the large size wires required for low output voltage switched mode power supplies. It also allows for a flow of air which keeps the assembly cooler. The center post is round, like that of the pot core. One of the advantages of the round center post is that the winding has a shorter path length around it (11% shorter) than the wire around a square center post with an equal area. This reduces the losses of the windings by 11% and enables the core to handle a higher output power. The round center post also eliminates the sharp bend in the wire that occurs with winding on a square center post.

PQ Cores

PQ cores are designed especially for switched mode power supplies. The design provides an optimized ratio of volume to winding area and surface area. As a result, both maximum inductance and winding area are possible with a minimum core size. The cores thus provide maximum power output with a minimum assembled transformer weight and volume, in addition to taking up a minimum amount of area on the printed circuit board. Assembly with printed circuit bobbins and one piece clamps is simplified. This efficient design provides a more uniform cross-sectional area; thus cores tend to operate with fewer hot spots than with other designs.

EP Cores

EP Cores are round center-post cubical shapes which enclose the coil completely except for the printed circuit board terminals. The particular shape minimizes the effect of air gaps formed at mating surfaces in the magnetic path and provides a larger volume ratio to total space used. Shielding is excellent.

Toroids

Toroids are economical to manufacture; hence, they are least costly of all comparable core shapes. Since no bobbin is required, accessory and assembly costs are nil. Winding is done on toroidal winding machines. Shielding is relatively good.

Summary

Ferrite geometries offer a wide selection in shapes and sizes. When choosing a core for power applications, parameters shown in Table 1 should be evaluated.

TABLE 1
FERRITE CORE COMPARATIVE GEOMETRY CONSIDERATIONS

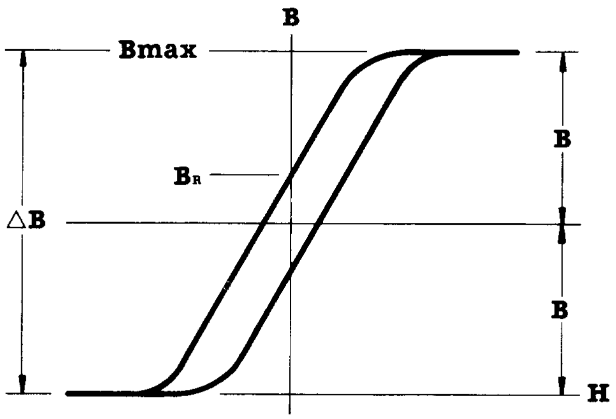
	Pot Core	Double Slab, RM Cores	E Core	EC, ETD, EER Cores	PQ Core	EP Core	Toroid
See Catalog Section	6	7-8	11	12	10	9	13
Core Cost	high	high	low	medium	high	medium	very low
Bobbin Cost	low	low	low	medium	high	high	none
Winding Cost	low	low	low	low	low	low	high
Winding Flexibility	good	good	excellent	excellent	good	good	fair
Assembly	simple	simple	simple	medium	simple	simple	none
Mounting Flexibility **	good	good	good	fair	fair	good	poor
Heat Dissipation	poor	good	excellent	good	good	poor	good
Shielding	excellent	good	poor	poor	fair	excellent	good

** Hardware is required for clamping core halves together and mounting assembled core on a circuit board or chassis.

TRANSFORMER CORE SIZE SELECTION

The power handling capacity of a transformer core can be determined by its WaAc product, where Wa is the available core window area, and Ac is the effective core cross-sectional area.

FIGURE 1



The WaAc/power-output relationship is obtained by starting with Faraday's Law:

$E = 4B\ Ac\ Nf \times 10^{-8}$ (square wave) (1)

$E = 4.44\ BAc\ Nf \times 10^{-8}$ (sine wave) (1a)

Where:

E = applied voltage (rms)	K = winding factor
B = flux density in gauss	I = current (rms)
Ac = core area in cm^2	P_i = input power
N = number of turns	P_o = output power
f = frequency in Hz	e = transformer efficiency
Aw = wire area in cm^2	
Wa = window area in cm^2: core window for toroids, bobbin window for other cores	
C = current capacity in cm^2/amp	

Solving (1) for NAc

$$NAc = \frac{E \times 10^8}{4Bf} \quad (2)$$

The winding factor

$$K = \frac{NAw}{Wa} \quad \text{thus } N = \frac{KWa}{Aw} \quad \text{and } NAc = \frac{KWaAc}{Aw} \quad (3)$$

Combining (2) and (3) and solving for WaAc:

$$WaAc = \frac{E\ Aw \times 10^8}{4B\ fK}, \text{ where } WaAc = cm^4 \quad (4)$$

In addition:

$C = Aw/I$ or $Aw = IC$ $e = P_o/P_i$ $P_i = EI$

Thus:

$E\ Aw = EIC = P_iC = P_oC/e$

Substituting for EAw in (4), we obtain:

$$WaAc = \frac{P_oC \times 10^8}{4eB\ fK} \quad (5)$$

Assuming the following operational conditions:

- C = 4.05 x $10^{-3}cm^2$/Amp (square wave) and 2.53 x $10^{-3}cm^2$/Amp (sine wave) for toroids
- C = 5.07 x $10^{-3}cm^2$/Amp (square wave) and 3.55 x $10^{-3}cm^2$/Amp (sine wave) for pot cores and E-U-I cores.
- e = 90% for transformers
- e = 80% for inverters (including circuit losses)
- K = .3 for pot cores and E-U-I cores (primary side only)
- K = .2 for toroids (primary side only)

With larger wire sizes, and/or higher voltages, these K factors may not be obtainable. To minimize both wire losses and core size, the window area must be full.

NOTE: For Wire Tables and turns/bobbin data, refer to pages 5.9, 5.10 and 5.11.

We obtain the basic relationship between output power and the WaAc product:

$$WaAc = \frac{k'\ Po \times 10^8}{Bf}, \text{ Where } k' = \frac{C}{4eK}$$

For square wave operation

k'=.00633 for toroids, k'=.00528 for pot cores, k'=.00528 for E-U-I cores

If we assume B to be 2000 gauss, we obtain the graphs on Figure 7, page 4.7 showing output power as a function of WaAc and frequency.

Cores selected using these graphs will generate the following temperature increases for 20 KHz **square wave** operation @ 2000 gauss (typical inverter output transformer conditions):

WaAc range	Δ T range
< 1.01	< 30°C
1.01 to 4.5	30°C to 60°C
4.5 to 15.2	60°C to 90°C

These temperatures can be reduced, as needed, by lowering the flux density or frequency, increasing the wire diameter, or using special cooling techniques.

Due to the wide range of core sizes and operational conditions, it is essential that your design be evaluated for temperature rise before the final core size is chosen.

CIRCUIT TYPES

Some general comments on the different circuits are:

The push-pull circuit is efficient because it makes bidirectional use of a transformer core, providing an output with low ripple. However, circuitry is more complex, and the transformer core saturation can cause transistor failure if power transistors have unequal switching characteristics.

Feed forward circuits are low in cost, using only one transistor. Ripple is low because relatively steady state current flows in the transformer whether the transistor is ON or OFF.

The flyback circuit is simple and inexpensive. In addition, EMI problems are less. However, the transformer is larger and ripple is higher.

TABLE 2 CIRCUIT TYPE SUMMARY

Circuit	Advantages	Disadvantages
Push-pull	Medium to high power Efficient core use Ripple and noise low	More components
Feed forward	Medium power Low cost Ripple and noise low	Core use inefficient
Flyback	Lowest cost Few components	Ripple and noise high Regulation poor Output power limited (< 100 watts)

PUSH-PULL CIRCUIT

A typical push-pull circuit is shown in Figure 2a. The input signal is the output of an IC network, or clock, which switches the transistors alternately ON and OFF. High frequency square waves on the transistor output are subsequently rectified, producing dc.

FIGURE 2a
TYPICAL PUSH-PULL SPS CIRCUIT

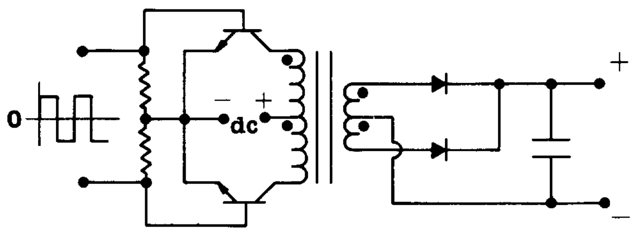
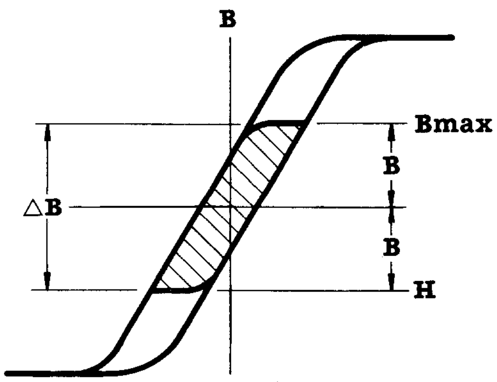


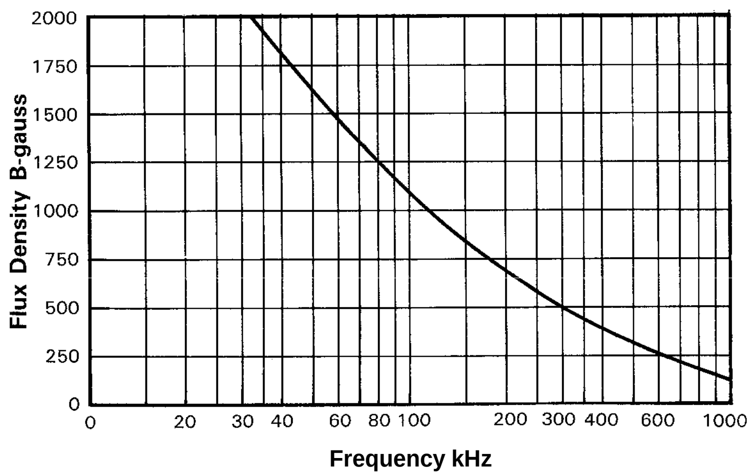
FIGURE 2b
HYSTERESIS LOOP OF MAGNETIC CORE IN PUSH-PULL CIRCUIT



For ferrite transformers, at 20 kHz, it is common practice to apply equation (4) using a flux density (B) level of ± 2 kG maximum. This is illustrated by the shaded area of the Hysteresis Loop in figure 2b. This B level is chosen because the limiting factor in selecting a core at this frequency is core loss. At 20 kHz, if the transformer is designed for a flux density close to saturation (as done for lower frequency designs), the core will develop an excessive temperature rise. Therefore, the lower operating flux density of 2 kG will usually limit the core losses, thus allowing a modest temperature rise in the core.

Above 20 kHz, core losses increase. To operate the SPS at higher frequencies, it is necessary to operate the core flux levels lower than ± 2 kg. Figure 3 shows the reduction in flux levels for MAGNETICS “P” ferrite material necessary to maintain constant 100mW/cm^3 core losses at various frequencies, with a maximum temperature rise of 25°C.

FIGURE 3



FEED FORWARD CIRCUIT

FIGURE 4a
TYPICAL FEED FORWARD SPS CIRCUIT

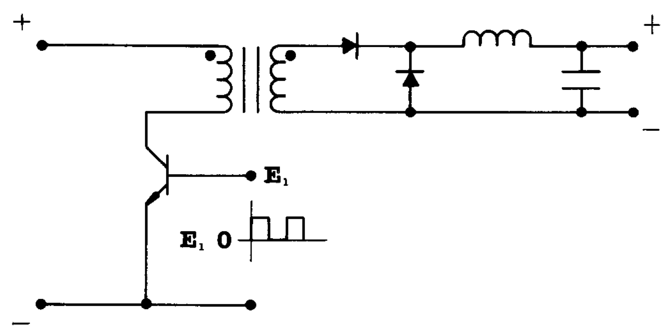
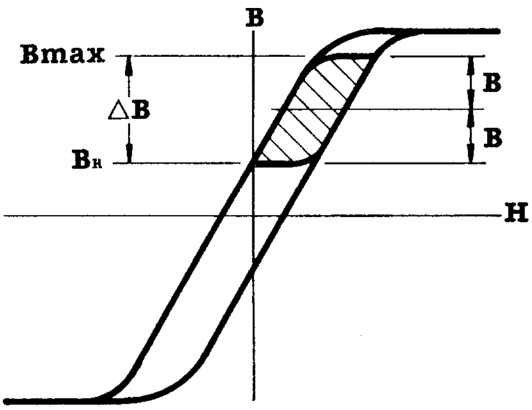


FIGURE 4b
HYSTERESIS LOOP OF
MAGNETIC CORE IN
FEED FORWARD CIRCUIT



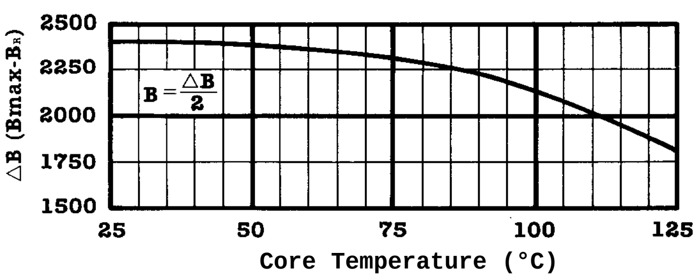
In the feed forward circuit shown in Figure 4a, the transformer operates in the first quadrant of the Hysteresis Loop. (Fig. 4b). Unipolar pulses applied to the semiconductor device cause the transformer core to be driven from its B_R value toward saturation. When the pulses are reduced to zero, the core returns to its B_R value. In order to maintain a high efficiency, the primary inductance is kept high to reduce magnetizing current and lower wire losses. This means the core should have a zero or minimal air gap.

For ferrites used in this circuit, Δ B (or B max-B_R) is typically 2400 gauss or B (as applied to Equation 4) is ± 1200 gauss as shown in Figure 4b. In the push-pull circuit, it was recommended that the peak flux density in the core should not exceed B = ± 2000 gauss in order to keep core losses small. Because of the constraints of the Hysteresis Loop, the core in the feed forward circuit should not exceed a peak value of B = ± 1200 gauss.

Core selection for a feed forward circuit is similar to the push-pull circuit except that B for Equation 4 is now limited to ± 1200 gauss. If the charts in Figure 7 are used, WaAc is selected from the appropriate graph and increased by the ratio of $\frac{2000}{1200}$ = 1.67, or by 67%.

If the transformer operating temperature is above 75°, the value of B will be further reduced. Figure 5 shows the variation of ΔB with temperature. Therefore, the recommended ΔB value of 2400 (B = ± 1200) gauss has to be reduced, the amount depending on the final projected temperature rise of the device.

FIGURE 5



The value of ΔB remains virtually unchanged over a large frequency range above 20 kHz. However, at some frequency, the adjusted value of B, as shown in Figure 3, will become less than the B determined by the above temperature considerations (Figure 5). Above this frequency, the B used to select a core will be the value obtained from Figure 3.

FLYBACK CIRCUIT

A typical schematic is shown in Figure 6a. Unipolar pulses cause dc to flow through the core winding, moving the flux in the core from B_R towards saturation (Fig. 6b). When the pulses go to zero, the flux travels back to B_R as in the feed forward design. However, the difference between the feed forward and the flyback circuit is that the flyback requires the transformer to act as an energy storage device as well as to perform the usual transformer functions. Therefore, to be an effective energy storage unit, the core must not saturate and is usually a gapped structure.

FIGURE 6a
TYPICAL FLYBACK REGULATOR
CIRCUIT

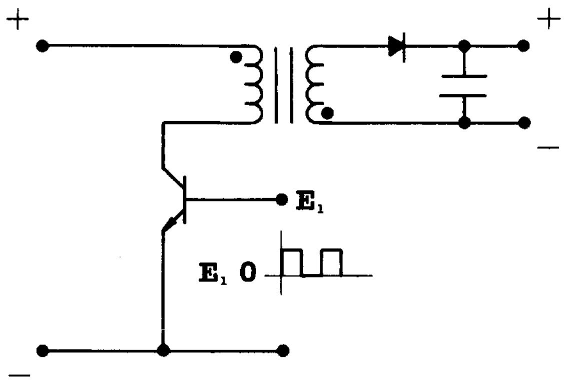
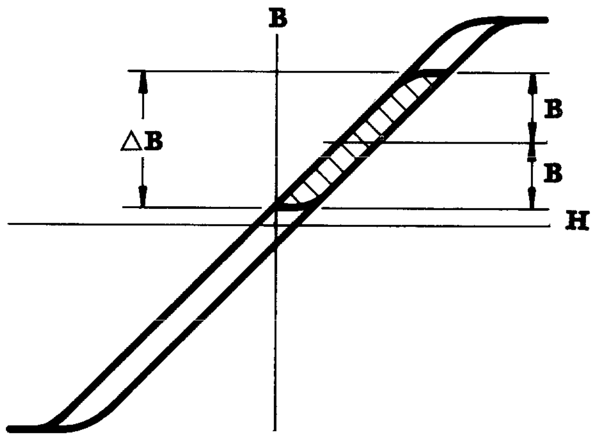


FIGURE 6b
HYSTERESIS LOOP OF MAGNETIC
CORE IN FLYBACK CIRCUIT



In most designs, the air gap is large; therefore, B_R is small as noted on the Hysteresis Loop in Figure 6b and can be considered zero. The maximum flux density available is approximately 3600. This means ΔB is 3600 or B = ± 1800 gauss. Core selection for this circuit can also be done using Equation 4 or the charts in Figure 7 as previously described. The B value in Equation 4 is ± 1800 gauss at 20 kHz and is used until a higher frequency (Figure 3) dictates a lower B is required.

To simplify core selection without using Equation (4) and under the limiting conditions noted above, graphs shown in Figure 7 can be used.

These graphs note the output power vs. core size at various frequencies. Core selection is made by determining the amount of output power required. Using one of the charts in Figure 7, find the intersection of the output power and the operating frequency line; the vertical projection of this point indicates the core to be used. If the vertical projection of the point is between two

cores, choose the larger one. If, for example, a ferrite pot core is needed for a transformer output power of 20 watts at 20 kHz, the above procedure indicates that the correct pot core is between a 42318-UG and a 42616-UG size. In this example, the large core (42616-UG) would be the better selection.
Above 20 kHz, this procedure is changed as follows. First note the intersection of the horizontal line representing the desired output power and the frequency of operation. The vertical line through this intersection intersects the horizontal axis which lists the appropriate WaAc. This factor is based on operating at B = ±2kG and must be increased inversely in proportion to the decreased flux density recommended for the operating frequency (see Figure 3). Using the newly selected WaAc factor on the graph, the vertical line through this point indicates the core to be used in this design.

In the example above (output of 20 watts), if the core is to operate at 50 kHz, the graph indicates a pot core with a WaAc of .117 cm^4. Figure 3, however, shows that the flux density at 50 kHz must be reduced to 1600 gauss. Therefore, the ratio

$$\frac{(WaAc)\ @\ 20kHz}{(WaAc)\ @\ 50kHz} \text{ equals } \frac{2000}{1600} \text{ or } 1.25.$$

The new WaAc at 50kHz must be .117 x 1.25 or .146 cm^4. From the graph, the larger WaAc value dictates a 42213-UG pot core.

GENERAL FORMULA — CORE SELECTION FOR DIFFERENT TOPOLOGIES

The following formula has been gained from derivations in Chapter 7 of A. I. Pressman's book, "Switching Power Supply Design" (see Reference No. 14, page 14.1).

$$WaAc = \frac{P_o D_{cma}}{K_t B_{max} f}$$

WaAc = Product of window area and core area (cm^4)
P_o = Power Out (watts)
D_{cma} = Current Density (cir. mils/amp)
B_{max} = Flux Density (gauss)
f = frequency (hertz)
K_t = Topology constant (for a space factor of 0.4):
Forward converter = .0005 Push-pull = .001
Half-bridge = .0014 Full-bridge = .0014
Flyback = .00033 (single winding)
Flyback = .00025 (multiple winding)

For individual cores, WaAc is listed in this catalog under "Magnetic Data."

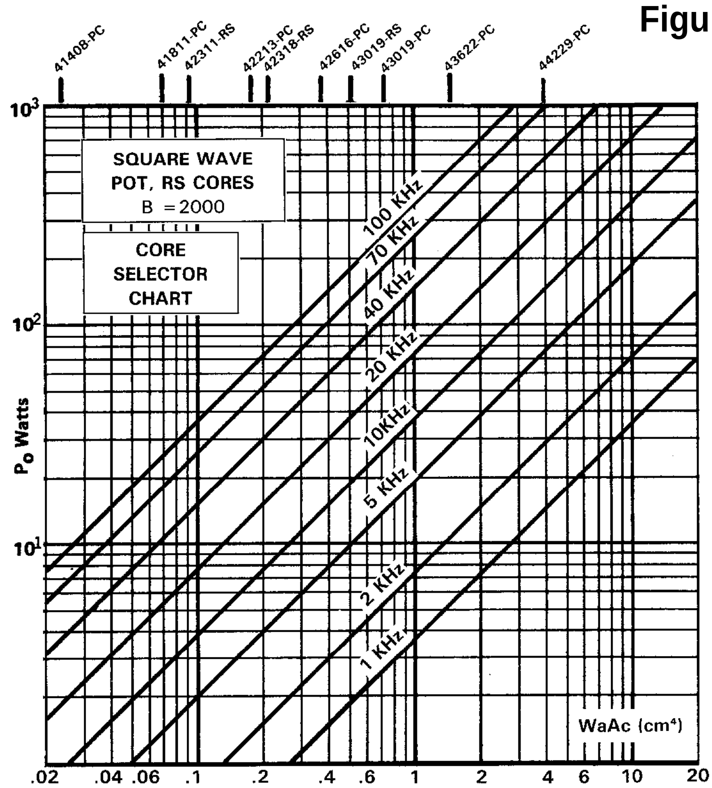
Choice of B_{max} at various frequencies, D_{cma} and alternative transformer temperature rise calculation schemes are also discussed in Chapter 7 of the Pressman book.

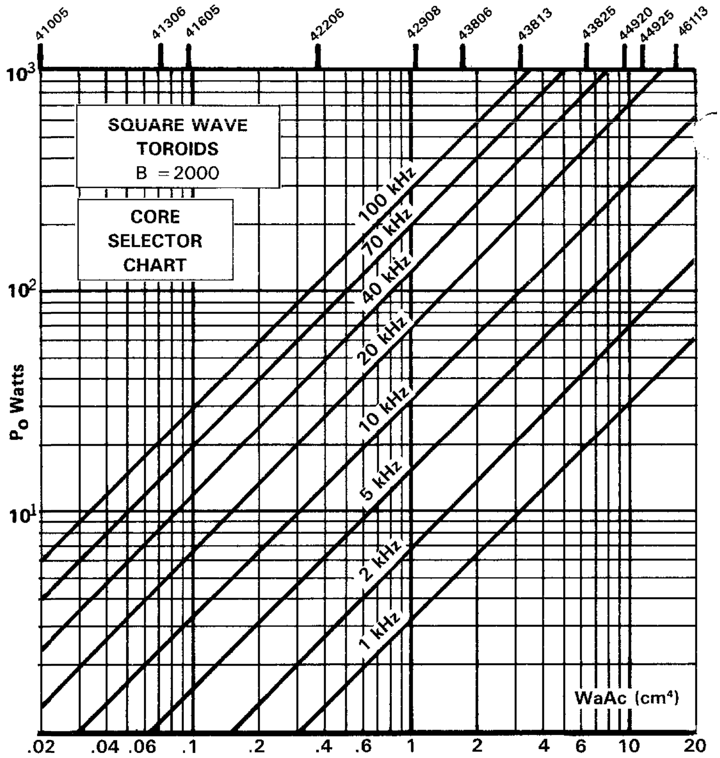
Table 3 - FERRITE CORE SELECTION BY AREA PRODUCT DISTRIBUTION

$W_a A_c$* (cm^4)	PC	RS, DS HS	RM, EP	RM SOLID	PQ	EE LAM	EE, EEM, EFD	EE, EI PLANAR	EC	ETD, EER	UU, UI	TC
.001	40704							41309(E)				40601
.002	40905		40707(EP)				40904 40906					40603
.004												
.007	41107		41110(RM)									40705
.010		41408 (RS,DS)	41010(EP)			41203					41106 (UI)	41003 41005
.020	41408		41510(RM) 41313(EP)	41510		41205	41208 41209 41515 41707				41106(U)	40907 41303
.040			41812(RM)	41812			41709 42110					41206 41305
.070	41811	42311 (RS, DS, HS)	41717(EP)		42610	41808						41306 41605
.100	42213	42318 (HS)	42316(RM)	42316	42016 42614	41810 42510		42216(E)				
.200	42616	42318 (RS, DS) 42616 (RS, DS, HS)	42819(RM) 42120(EP)		42020 42620 43214		42211 42810 43009 42523	43618(EI) 43208(EI)			42515 (UI)	41809 42206
.400		43019 (RS, DS, HS)		42819	42625	42520	42515 43007	43618(E) 43208(E)				42207
.700	43019		43723(RM)		43220	43515	43013		43517		42220(U) 42512(U) 42515(U)	42507
1.00	43622	43622 (RS, DS, HS)		43723	43230	44317	43520 43524 44011	44308(EI)	44119	43434 43521(EER)	42530(U)	42908
2.00	44229 44529	44229 (RS, DS, HS)			43535	44721	44020 44924	44308(E)	45224	43939 44216(EER) 44444	44119(U) 44121(U)	43610 43615 43813
4.00					44040	45724	44022 45021			44949	44125(U) 44130(U)	44416
7.00							45528 46016	46410(E)				
10.00							45530 47228		47035			44916 44925 46113
20.00							48020			47054		47313
40.00												48613
100											49925(U)	

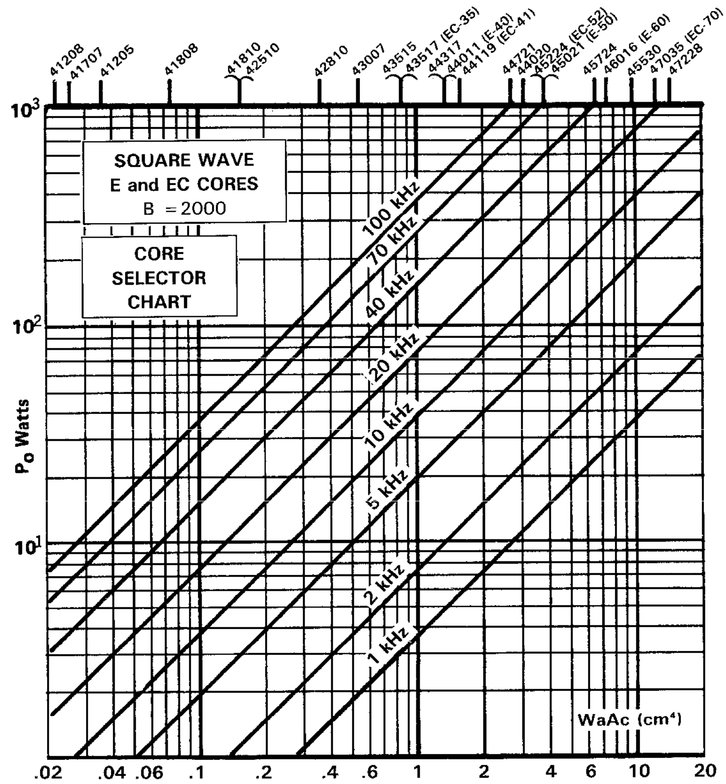
* Bobbin window & core area product. For bobbins other than those in this catalog, $W_a A_c$ may need to be recalculated.

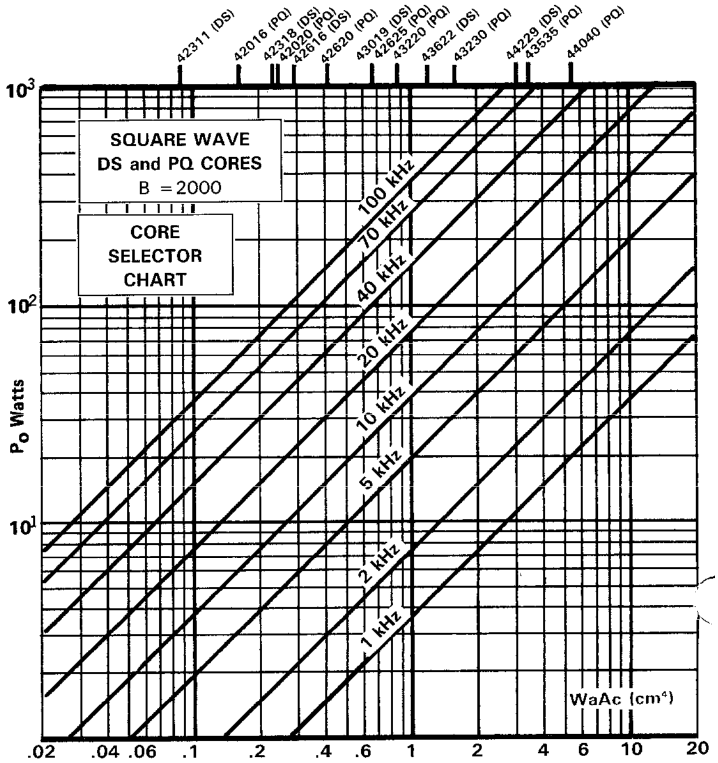
Figure 7





For a given P_O choose a core size which has a WaAc equal to or greater than the value obtained from the graph. If the frequency is variable, use the lowest operating frequency. *These curves are based on bi-polar operations.*	The power handling capacity of a given core is directly proportional to the flux level (B), the frequency (f), and the current density (1/C); thus the WaAc needed for your operational conditions can be readily obtained using these graphs. Note: These curves are based on B = 2000. At frequencies of 40 to 100 kHz, it may be more desirable to use B = 1000. If so, double the WaAc required.	For saturating square wave inverters, divide WaAc by 1.8 for F or P materials. G, J, & W materials are not recommended for this type application. Saturating inverters with P_O values > 50 watts or frequencies > 10 kHz are not recommended due to excessively high Δ T values.	For center tap transformers, the wire size can be reduced 30% (one wire size), but the WaAc values should be increased by 20% if the primary is push-pull, and 40% if both primary and secondary are push-pull.





TEMPERATURE CONSIDERATIONS

The power handling capability of ferrite cores is limited by the temperature rise of the transformer. Major factors affecting this temperature rise include wire loss, core loss, and core geometry.

Wire Loss - Wire loss with dc current increases with temperature (°C) as shown:

$$R_t = R_{25}\,[1 + .004(T-25)]$$

Where R_t = wire resistance at operational temperature
R_{25} = resistance at room temperature (25°C)
T = operating temperature (°C)

Skin effect losses and proximity losses become a significant factor at high frequencies. At 20 kHz (square wave), wire sizes larger than #24 will increase wire losses significantly. For high current/high frequency operation stranded wire, litz wire, strip winding, or other special windings are recommended.

Core Loss — Core loss versus flux density and frequency is shown graphically for each power material in section 3. A core loss of 100 mW/cm³ will result in an approximate temperature rise of 40°C. If the core loss due to the chosen flux density and/or frequency exceeds 100 mW/cm³, special cooling precautions (heat sinks and/or fans) may be necessary.

For high power operation, the power handling capability is proportional to the temperature rise (e.g., a unit with a temperature rise of 80°C yields 1.5 to 2 times the power output of the same unit with a temperature rise of only 40°C); thus ΔT should be as high as possible. Transformer hot spot temperatures, however, should not exceed 120°C.

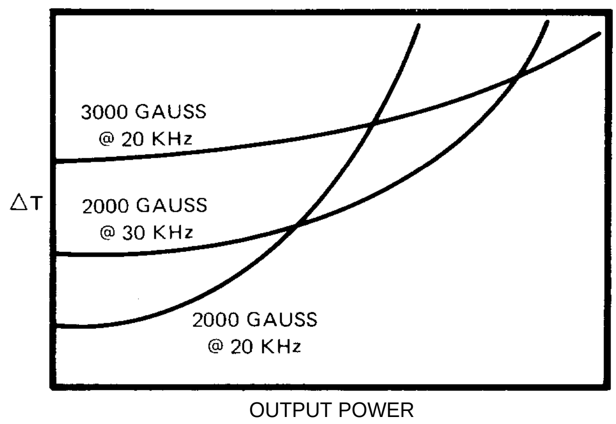
Cores driven with square waves exhibit slightly lower core losses than when driven with the sine waves; however, wire losses are much greater with square waves than with sine waves.

Core Geometry — Transformer temperature rise is directly related to the ratio of transformer surface area to transformer volume; larger cores are thus less efficient in radiating heat losses. For small cores (WaAc < .5) the graphical P_O will generate temperature rises <40°C; for large cores (WaAc > 1.0) the graphical P_O will generate temperature rises >60°C.

E-U-I cores offer optimum cooling geometry; toroids are satisfactory; pot cores are least effective as the coil is shielded by the core.

The winding area/core area ratio (Wa/Ac) is only .4 to .5 for pot cores, and .6 to 1.0 for E cores, and ranges from 1 to 10 for toroids. Thus, for a given WaAc, toroids will accept more wire, but will also saturate more readily than E cores or pot cores.

Typical output power versus temperature rise behavior (for any geometry) as a function of square wave frequency and flux density is shown in the following graph.



TRANSFORMER EQUATIONS

Once a core is chosen, the calculation of primary and secondary turns and wire size is readily accomplished.

$$N_p = \frac{V_p \times 10^8}{4BAf} \qquad N_s = \frac{V_s}{V_p} N_p$$

$$I_p = \frac{P_{in}}{E_{in}} = \frac{P_{out}}{eE_{in}} \qquad I_s = \frac{P_{out}}{E_{out}}$$

$$KWa = N_p A_{wp} + N_s A_{ws}$$

Where

A_{wp} = primary wire area A_{ws} = secondary wire area

Assume K = .4 for toroids; .6 for pot cores and E-U-I cores

Assume $N_p A_{wp} = 1.1\, N_s A_{ws}$ to allow for losses and fee? back winding

$$\text{efficiency } e = \frac{P_{out}}{P_{in}} = \frac{P_{out}}{P_{out} + \text{wire losses} + \text{core losses}}$$

$$\text{Voltage Regulation (\%)} = \frac{R_s + (N_s/N_p)^2 R_p}{R_{load}} \times 100$$

Table 4

FERRITE CORE SELECTION LISTED BY TYPICAL POWER HANDLING CAPACITIES (WATTS)
(F, P and R Materials) (For push-pull square wave operation, see notes below).

Wattage				E-Cores	EC-ETD, U Cores	PQ Cores	DS Cores	RS-RM-PC Cores	TC Toroids	EP Cores	Low Profile, lanar Cores
@f= 20 kHz	@f= 50 kHz	@f= 100 kHz	@f= 250 kHz								
2	3	4	7	41707				41408-PC	41206 41303	41313	41709 42107 42110
5	8	11	21	41808			42311	41811-PC 42311-RS,42809-RM	41306 41605	41717	42610-PQ 42216-EC
12	18	27	53	41810, 42211		42016		42316-RM			42614-PQ
13	20	30	59	42510							
15	22	32	62					42213-PC			
18	28	43	84			42020	42318	42318-RS	42106		43618-E, I
19	30	48	94				42616		41809	42120	43208-E, I 44008-E, I
26	42	58	113	42810,42520					42206		
28	45	65	127	42515				42819-RM	42109		
30	49	70	137			42620		42616-PC	42207		
33	53	80	156				43019				43618-EC
40	61	95	185	43007				43019-RS	43205		44008-EC
42	70	100	195			42625					43208-EC
48	75	110	215	43013					42212,42507		
60	100	150	293	42530, 43009 43515 (E375)	43517(EC35)	43220		43019-PC 43723-RM			
70	110	170	332		43434(ETD34)		43622		42908		44308-E,I
105	160	235	460	44011(E40)							
110	190	250	480			43230		43622-PC			
120	195	270	525		44119 (EC41)						
130	205	290	570	43524,43520	43521				43806		
140	215	340	663	44317(E21)					42915, 43113		
150	240	380	741		43939 (ETD39)						44308-EC
190	300	470	917				44229		43610		
200	310	500	975	44721(E625)							
220	350	530	1034			43535			43813		
230	350	550	1073	44020(42/15)	44216						
260	400	600	1170						43615		
280	430	650	1268	45021(E50), 44924	45224(EC52)			44229-PC			
300	450	700	1365	44022(42/20)	44444(ETD44)			44529-PC			
340	550	850	1658			44040					
360	580	870	1697						43825		
410	650	1000	1950	45724(E75)	44949(ETD49)				44416		
550	800	1300	2535	45528(55/21) 46016(E60)					44715		
650	1000	1600	3120						44916,44920		46410-EC
700	1100	1800	3510	45530(55/25)							
850	1300	1900	3705						44925		
900	1500	2000	3900		47035 (EC70)						
1000	1600	2500	4875		45959 (ETD59)				46113		
1000	1700	2700	5265	47228							
1400	2500	3200	6240						44932		
1600	2600	3700	7215						47313		
2000	3000	4600	8970	48020	47054						
2800	4200	6500	12675						48613		
11700	19000	26500	51500		49925 (U)						

Above is for push-pull converter. De-rate by a factor of 3 or 4 for flyback. De-rate by a factor of 2 for feed-forward converter.

Note: Assuming Core Loss to be Approximately 100mW/cm³,
B Levels Used in this Chart are:
@ 20 kHz - 2000 gauss @ 50 kHz - 1300 gauss @ 100 kHz - 900 gauss @ 250 kHz - 700 gauss

See page 4.6 - Area Product Distribution

EMI FILTERS

Switch Mode Power Supplies (SMPS) normally generate excessive high frequency noise which can affect electronic equipment like computers, instruments and motor controls connected to these same power lines. An EMI Noise Filter inserted between the power line and the SMPS eliminates this type of interference (Figure 8). A differential noise filter and a common mode noise filter can be in series, or in many cases, the common mode filter is used alone.

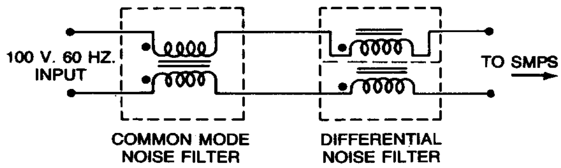


Figure 8

Common Mode Filter

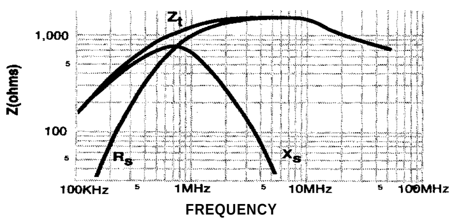
In a CMN filter, each winding of the inductor is connected in series with one of the input power lines. The connections and phasing of the inductor windings are such that flux created by one winding cancels the flux of the second winding. The insertion impedance of the inductor to the input power line is thus zero, except for small losses in the leakage reactance and the dc resistance of the windings. Because of the opposing fluxes, the input current needed to power the SMPS therefore will pass through the filter without any appreciable power loss.

Common mode noise is defined as unwanted high frequency current that appears in one or both input power lines and returns to the noise source through the ground of the inductor. This current sees the full impedance of either one or both windings of the CMN inductor because it is not canceled by a return current. Common mode noise voltages are thus attenuated in the windings of the inductor, keeping the input power lines free from the unwanted noise.

Choosing the Inductor Material

A SMPS normally operates above 20kHz. Unwanted noises generated in these supplies are at frequencies higher than 20kHz, often between 100kHz and 50MHz. The most appropriate and cost effective ferrite for the inductor is one offering the highest impedance in the frequency band of the unwanted noise. Identifying this material is difficult when viewing common parameters such as permeability and loss factor. Figure 9 shows a graph of impedance Z_t vs. frequency for a ferrite toroid, J-42206-TC wound with 10 turns.

Figure 9



The wound unit reaches its highest impedance between 1 and 10MHz. The series inductive reactance X_s and series resistance R_s (functions of the permeability and loss factor o the material) together generate the total impedance Z_t.

Figure 10 shows permeability and loss factor of the ferrite material in Figure 9 as a function of frequency.The falling off of permeability above 750kHz causes the inductive reactance to fall. Loss factor, increasing with frequency, causes the resistance to dominate the source of impedance at high frequencies.

Figure 10

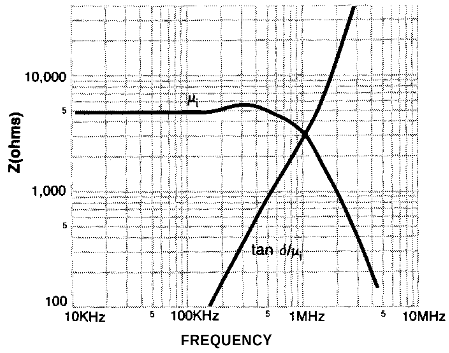
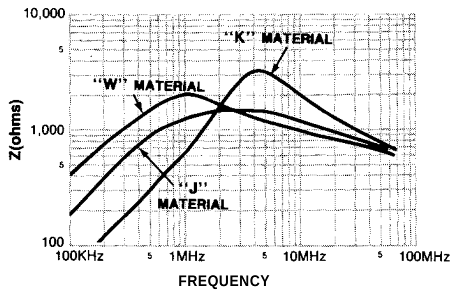


Figure 11 shows total impedance vs. frequency for three different materials. J material has a high total impedance over the range of 1 to 20MHz. It is most widely used for common mode filter chokes. Under 1Mhz, W material has 20-50% more impedance than J. It is often used in place of J when low frequency noise is the major problem. K material can be used above 2MHz because it produces up to 100% more impedance than J in this frequency range. For filter requirements specified at frequencies above and below 2MHz, either J or W is preferred.

Figure 11



Core Shape

Toroids are most popular for a CMN filter as they are inexpensive and have low leakage flux. A toroid must be wound by hand (or individually on a toroid winding machine). Normally a non-metallic divider is placed between the two windings, and the wound unit is epoxied to a printed circuit header for attaching to a pc board.

An E core with its accessories is more expensive than a toroid, but assembly into a finished unit is less costly. Winding E core bobbins is relatively inexpensive. Bobbins with dividers for separating the two windings are available for pc board mounting.

E cores have more leakage inductance, useful for differential filtering in a common mode filter. E cores can be gapped to increase the leakage inductance, providing a unit that will absorb both the common mode and differential unwanted noise.

Additional detailed brochures and inductor design software for this application are available from Magnetics.

Core Selection

The following is a design procedure for a toroidal, single-layer common mode inductor, see Figure 12. To minimize winding capacitance and prevent core saturation due to asymmetrical windings, a single layer design is often used. This procedure assumes a minimum of thirty degrees of free spacing between the two opposing windings.

The basic parameters needed for common mode inductor design are current (I), impedance (Z_S), and frequency (f). The current determines the wire size. A conservative current density of 400 amps/cm² does not significantly heat up the wire. A more aggressive 800 amps/cm² may cause the wire to run hot. Selection graphs for both levels are presented.

The impedance of the inductor is normally specified as a minimum at a given frequency. This frequency is usually low enough to allow the assumption that the inductive reactance, X_S, provides the impedance, see Figure 9. Subsequently, the inductance, L_S, can be calculated from:

$$L_S = \frac{X_S}{2\pi f} \qquad (1)$$

With the inductance and current known, Figures 13 and 14 can be used to select a core size based on the LI product, where L is the inductance in mH and I is the current in amps. The wire size (AWG) is then calculated using the following equation based on the current density (C_d) of 400 or 800 amps/cm²:

$$AWG = -4.31 \times \ln\left(\frac{1.889\, I}{C_d}\right) \qquad (2)$$

The number of turns is determined from the core's A_L value as follows:

$$N = \left(\frac{L_S \times 10^6}{A_L}\right)^{1/2} \qquad (3)$$

Design Example

An impedance of 100Ω is required at 10kHz with a current of 3 amps. Calculating the inductance from equation 1, L_S = 1.59 mH.

With an LI product of 4.77 at 800 amps/cm², Figure 14 yields the core size for chosen material. In this example, W material is selected to give high impedance up to 1MHZ, see Figure 11. Figure 14 yields the core W-41809-TC. Page 13.2 lists the core sizes and A_L values. Using an A_L of 12,200 mH/1000 turns, equation 3 yields N = 12 turns per side. Using 800 amps/cm², equation 2 yields AWG = 21.

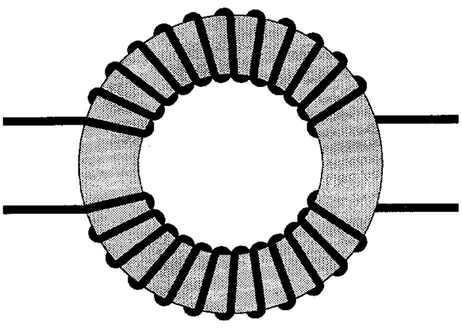


Figure 12: Common mode inductor winding arrangement

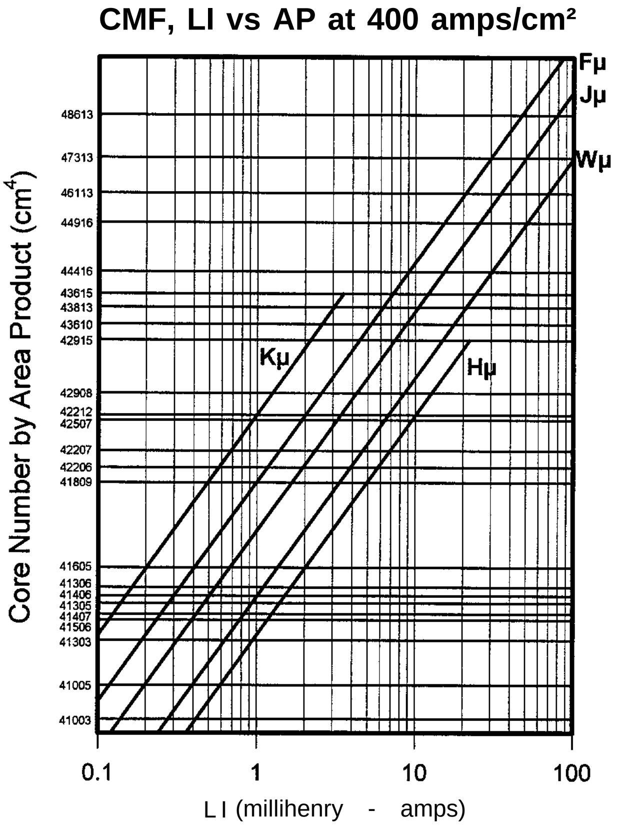


Figure 13: Core selection at 400 amps/cm²

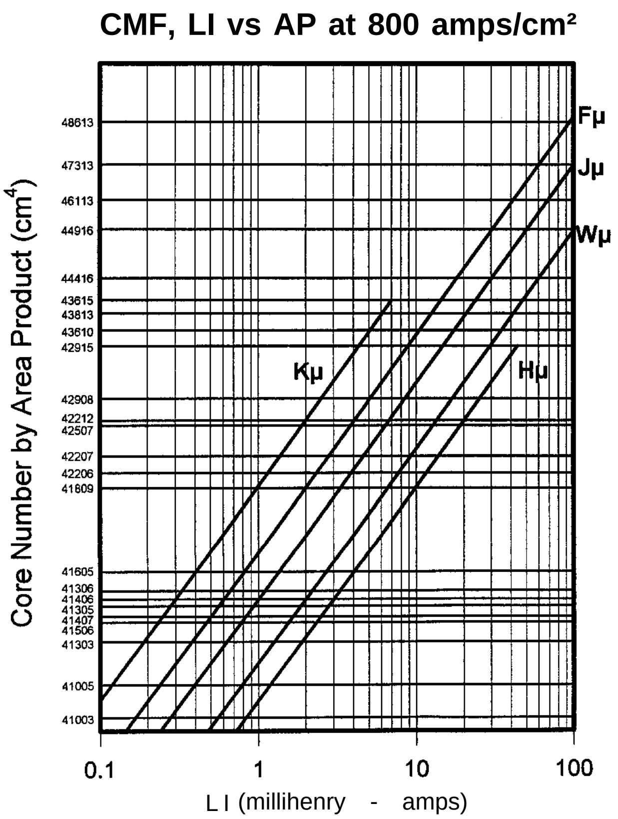


Figure 14: Core selection at 800 amps/cm²

HALL EFFECT DEVICES

Edwin H. Hall observed the "Hall Effect" phenomenon at Johns Hopkins University in 1897. He monitored the current flowing from top to bottom in a thin rectangular strip of gold foil by measuring the voltages at the geometric center of the left edge and the right edge of the strip. When no magnetic field was present, the voltages were identical. When a magnetic field was present perpendicular to the strip, there was a small voltage difference of a predictable polarity and magnitude. The creation of the transverse electric field, which is perpendicular to both the magnetic field and the current flow, is called the Hall Effect or Hall Voltage.

In metals the effect is small, but in semiconductors, considerable Hall voltages can be developed. Designers should consider using Hall sensors in many applications where mechanical or optical sensors have traditionally been used. To monitor ac or dc current flowing in a wire, the wire is wrapped around a slotted ferromagnetic core, creating an electro-magnet. The strength of the resulting magnetic field is used by the Hall sensor, inserted in the air gap, to measure the magnitude and direction of current flowing in the wire.

Core Selection

In all cases, the effective permeability of a gapped core will be a function of the size of the air gap and the initial permeability of the core material. Once the gap becomes greater than a few thousandths of an inch, the effective permeability is determined essentially by the air gap.

Analytical Method

1. Determine the flux operating extremes based on either the $\Delta V/\Delta B$ of the circuit (volts/gauss), or the maximum flux sensitivity (gauss) of the sensor (as provided by the sensor data sheet).

2. Choose a core based on the maximum or minimum dimension requirements to allow windings, and based on the core cross-section dimensions. The cross-section dimensions should be at least twice the gap length to ensure a relatively homogeneous flux distribution bridging the gap.

3. Calculate the maximum required μ_e for the core:

$$\mu_e = \frac{B\,l_e}{.4\pi NI} \qquad (1)$$

where: B = flux density (gauss)
l_e = path length (cm)
N = turns
I = current (amps peak)

4. Calculate the minimum required gap length (inches):

$$l_g = l_e\left(\frac{1}{\mu_e} - \frac{1}{\mu_i}\right)(0.3937) \qquad (2)$$

where: l_g = gap length (inches)
l_e = path length (cm)
μ_e = effective permeability
μ_i = initial permeability

5. If the minimum required gap is greater than the sensor thickness, ensure that the cross-section dimensions (length and width) are at least twice the gap length. If not, choose a larger core and recalculate the new gap length.

Graphical Method

1. Calculate NI/B (amp turns per gauss), knowing the flux operating extremes of $\Delta V/\Delta B$ or the maximum B sensitivity of the sensor.

2. Using Figure 15, follow the NI/B value from the vertical axis to the diagonal line to choose a ferrite core size. Drop down from the diagonal line to the horizontal axis to determine the gap length. The core sizes indicated on the selector chart take into account gap length versus cross-section dimensions in order to maintain an even flux distribution across the gap under maximum current.

Toroid Gapping

Ferrite cores are a ferromagnetic ceramic material. As such, they exhibit a very high hardness characteristic, they are very brittle, and they do not conduct heat very efficiently. Machining a slot into one side of a ferrite toroid can be a difficult process. Special techniques must be used to prevent chipping, cracking, or breaking of the cores.

Diamond bonded-tool machining is the preferred method of cutting ferrite. The bonded diamond particle size should be approximately 100 to 170 mesh (150 to 90 µm). The peripheral speed of the cutting wheel should be 5000 to 6000 feet/minute (1500 to 1800 meters/minute). The depth of the cut may be as deep as 1" (25 mm), but in order to minimize residual stress, the cut should be limited to a maximum of 0.250" (6 mm) per pass, the smaller the better. During all cutting, the wheel and core should be flooded with ample amounts of coolant water to provide a lubricant as well as remove heat buildup that would cause thermal stress cracking of the core.

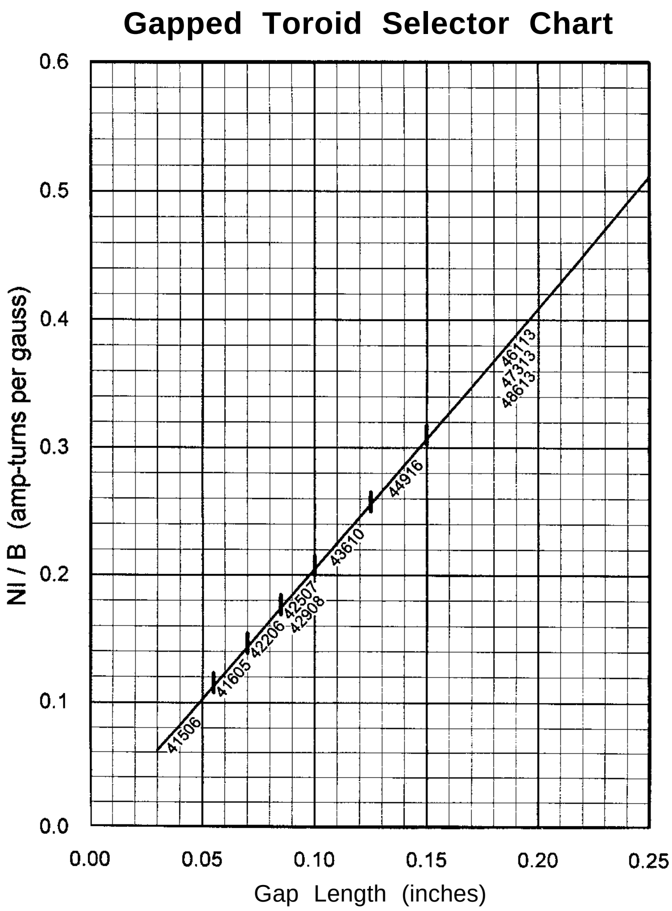
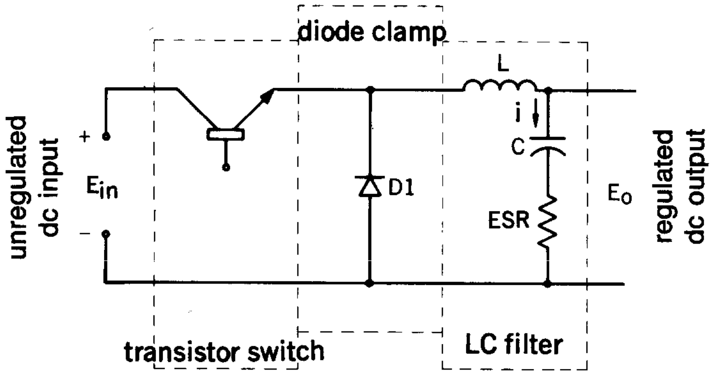


Figure 15: Hall Effect Device, Core Selector Chart

INDUCTOR CORE SIZE SELECTION (using core selector charts)

DESCRIPTION



A typical regulator circuit consists of three parts: transistor switch, diode clamp, and an LC filter. An unregulated dc voltage is applied to the transistor switch which usually operates at a frequency of 1 to 50 kilohertz. When the switch is ON, the input voltage, E_{in}, is applied to the LC filter, thus causing current through the inductor to increase; excess energy is stored in the inductor and capacitor to maintain output power during the OFF time of the switch. Regulation is obtained by adjusting the ON time, t_{on}, of the transistor switch, using a feedback system from the output. The result is a regulated dc output, expressed as:

$$E_{out} = E_{in}\, t_{on}\, f \qquad (1)$$

COMPONENT SELECTION

The switching system consists of a transistor and a feedback from the output of the regulator. Transistor selection involves two factors — (1) voltage ratings should be greater than the maximum input voltage, and (2) the frequency cut-off characteristics must be high compared to the actual switching frequency to insure efficient operation. The feedback circuits usually include operational amplifiers and comparators. Requirements for the diode clamp are identical to those of the transistor. The design of the LC filter stage is easily achieved. Given (1) maximum and minimum input voltage, (2) required output, (3) maximum allowable ripple voltage, (4) maximum and minimum load currents, and (5) the desired switching frequency, the values for the inductance and capacitance can be obtained. First, off-time (t_{off}) of the transistor is calculated.

$$t_{off} = (1 - E_{out}/E_{in\ max})/f \qquad (2)$$

When E_{in} decreases to its minimum value,

$$f_{min} = (1 - E_{out}/E_{in\ min})/t_{off} \qquad (3)$$

With these values, the required L and C can be calculated.

Allowing the peak to peak ripple current (Δi) through the inductor to be given by

$$\Delta i = 2\, I_{o\ min} \qquad (4)$$

the inductance is calculated using

$$L = E_{out}\, t_{off} / \Delta i \qquad (5)$$

The value calculated for Δi is somewhat arbitrary and can be adjusted to obtain a practical value for the inductance.
The minimum capacitance is given by

$$C = \Delta i / 8 f_{min} \Delta e_o \qquad (6)$$

Finally, the maximum ESR of the capacitor is

$$ESR\ max = \Delta e_0 / \Delta i \qquad (7)$$

INDUCTOR DESIGN

Ferrite E cores and pot cores offer the advantages of decreased cost and low core losses at high frequencies. For switching regulators, F or P materials are recommended because of their temperature and dc bias characteristics. By adding air gaps to these ferrite shapes, the cores can be used efficiently while avoiding saturation.

These core selection procedures simplify the design of inductors for switching regulator applications. One can determine the smallest core size, assuming a winding factor of 50% and wire current carrying capacity of 500 circular mils per ampere.

Only two parameters of the design application must be known:
(a) Inductance required with dc bias
(b) dc current.

1. Compute the product of LI^2 where:
   L = inductance required with dc bias (millihenries)
   I = maximum dc output current = $I_{o\ max} + \Delta i$

2. Locate the LI^2 value on the Ferrite Core Selector charts on pages 4.15 to 4.18. Follow this coordinate in the intersection with the first core size curve. Read the maximum nominal inductance, A_L, on the Y-axis. This represents the smallest core size and maximum A_L at which saturation will be avoided.

3. Any core size line that intersects the LI^2 coordinate represents a workable core for the inductor if the core's A_L value is less than the maximum value obtained on the chart.

4. Required inductance L, core size, and core nominal inductance (A_L) are known. Calculate the number of turns using

$$N = 10^3 \sqrt{\frac{L}{A_L}}$$

where L is in millihenries.

5. Choose the wire size from the wire table on page 5.9 using 500 circular mils per amp.

Example.

Choose a core for a switching regulator with the following requirements:

E_O	= 5 volts
Δe_o	= .5 volts
$I_{o\ max}$	= 6 amps
$I_{o\ min}$	= 1 amp
$E_{in\ min}$	= 25 volts
$E_{in\ max}$	= 35 volts
f	= 20 KHz

1. Calculate the off-time and minimum switching, f_{min}, of the transistor switch using equations 2 and 3.

$t_{off} = (1 - 5/35)/20{,}000 = 4.3 \times 10^{-5}$ seconds and

$f_{min} = (1 - 5/25)/4.3 \times 10^{-5}$ seconds $= 18{,}700$ Hz.

2. Let the maximum ripple current, Δi, through the inductor be

$\Delta i = 2(1) = 2$ amperes by equation 4.

3. Calculate L using equation 5.

$L = 5\ (4.3 \times 10^{-5})\ /2 = .107$ millihenries

4. Calculate C and ESR max using equations 6 and 7.

$C = 2/8\ (18700)\ (.5) = 26.7\ \mu$farads

and ESR max = .5/2 = .25 ohms

5. The product of $LI^2 = (.107)\ (8)^2 = 6.9$ millijoules.

6. Due to the many shapes available in ferrites, there can be several choices for the selecton. Any core size that the LI^2 coordinate intersects can be used if the maximum A_L is not exceeded. Following the LI^2 coordinate, the choices are:

(a) 45224 EC 52 core, A_L 315
(b) 44229 solid center post core, A_L 315
(c) 43622 pot core, A_L 400
(d) 43230 PQ core, A_L 250

7. Given the A_L, the number of turns needed for the required inductance is:

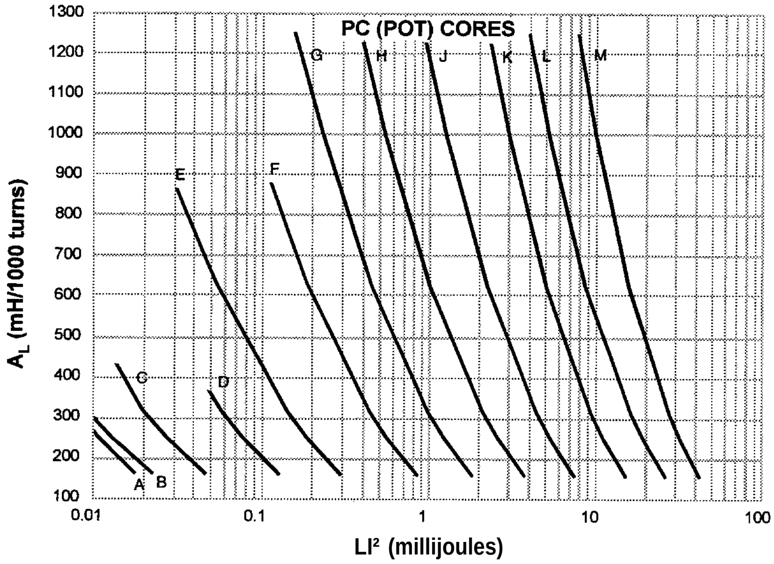
A_L	Turns
250	21
315	19
400	17

8. Use #14 wire.

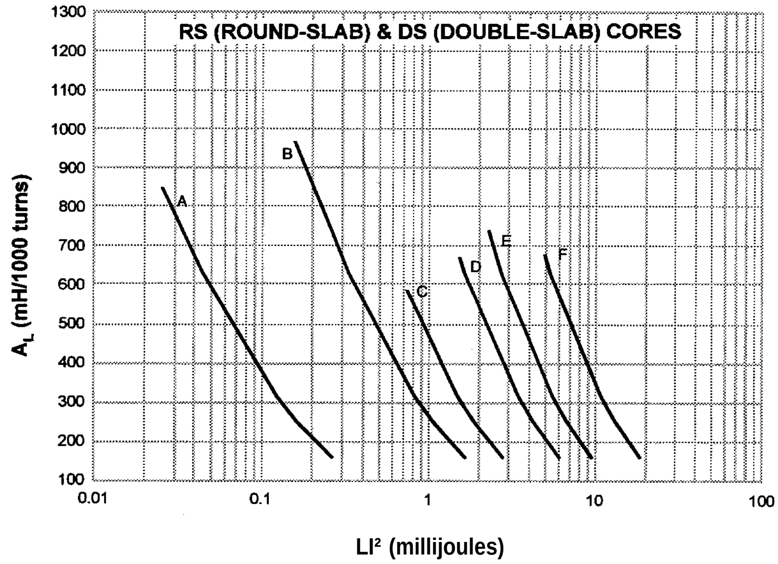
Note: MAGNETICS® Molypermalloy and Kool Mu® powder cores have a distributed air gap structure, making them ideal for switching regulator applications. Their dc bias characteristics allow them to be used at high drive levels without saturating. Information is available in catalogs MPP-303, KMC-01 and Brochure SR-IA.

FOR REFERENCES, See page 14.1.

Ferrite DC Bias Core Selector Charts



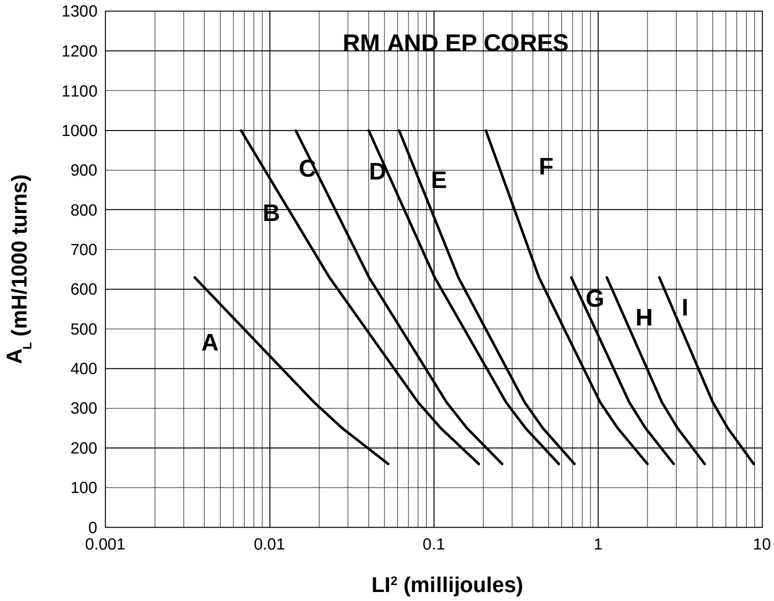
A- 40903
B- 40704
C- 40905
D- 41107
E- 41408
F- 41811
G- 42213
H- 42616
J- 43019
K- 43622
L- 44229
M- 44529



A- 41408 (RS)

B- 42311 (DS, RS)
42318 (DS, RS)

C- 42616 (DS)
D- 43019 (DS, RS)
E- 43622 (DS)
F- 44229 (DS)



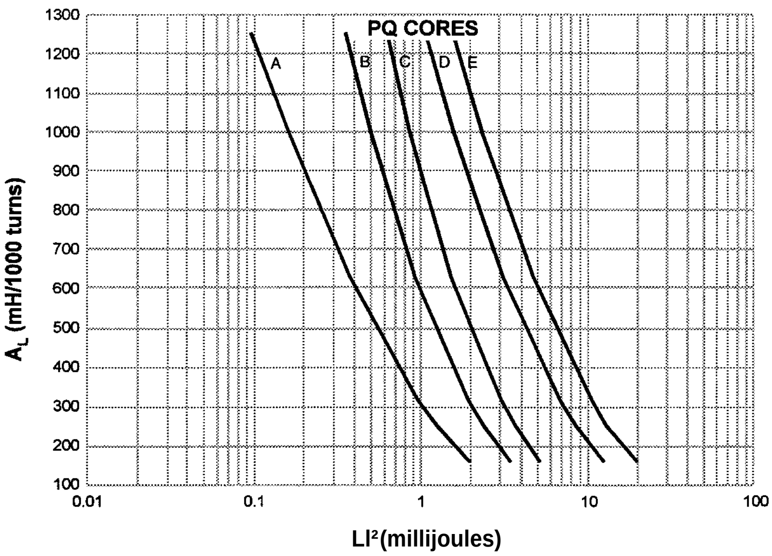
A- 40707 (EP7)
41010 (EP10)
41110 (RM4)

B- 41313 (EP13)
C- 41510 (RM5)
D- 41717 (EP17)
E- 41812 (RM6)
F- 42316 (RM8)
G- 42120 (EP20)

H- 42809 (RM10 planar)
42819 (RM10)

J- N43723 (RM12)

Ferrite DC Bias Core Selector Charts



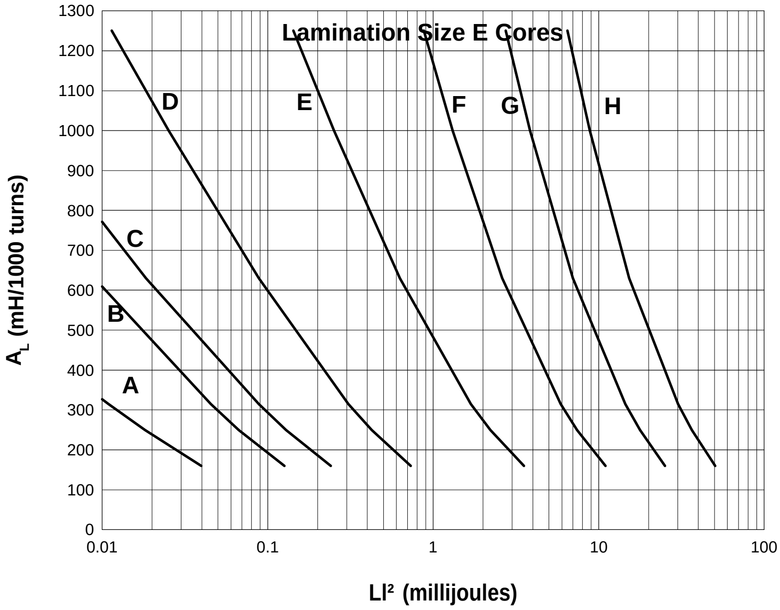
A- 42016
42020

B- 42614

C- 42610
42620
42625
43214

D- 43220
43230

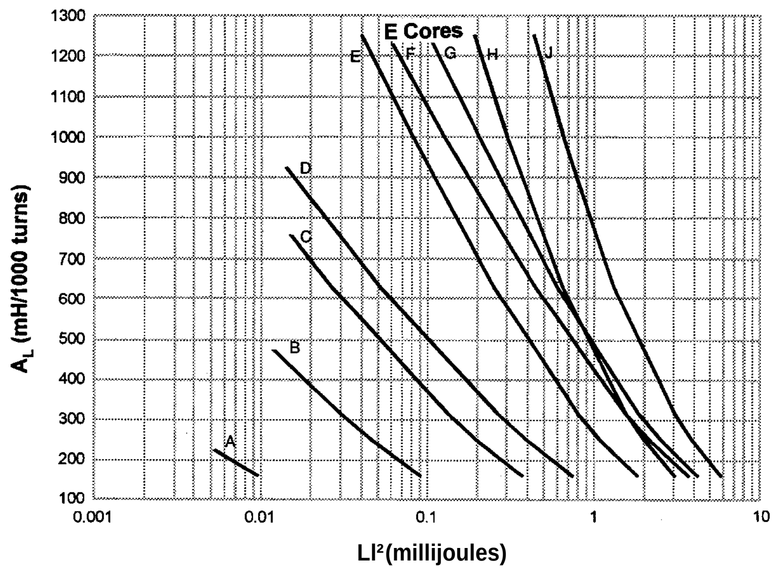
E- 43535
44040



A- 41203 (EE)
B- 41707 (EE)
C- 41808 (EE)
D- 42510 (EE)

E- 43009 (EE)
43515 (EE)

F- 44317 (EE)
G- 44721 (EE)
H- 45724 (EE)



A- 40904 (EE)

B- 41208 (EE)
41209 (EE)

C- 41205 (EE)
42211 (EE)

D- 42515 (EE)

E- 41810 (EE)
43007 (EE)

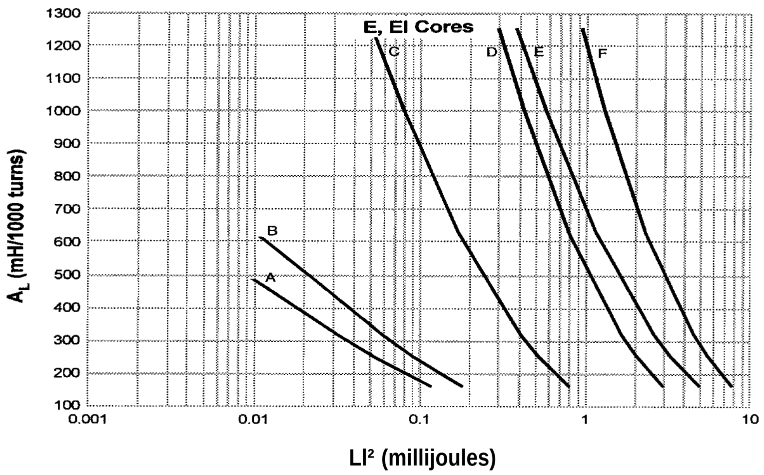
F- 43524 (EE)

G- 42530 (EE)
43520 (EE)

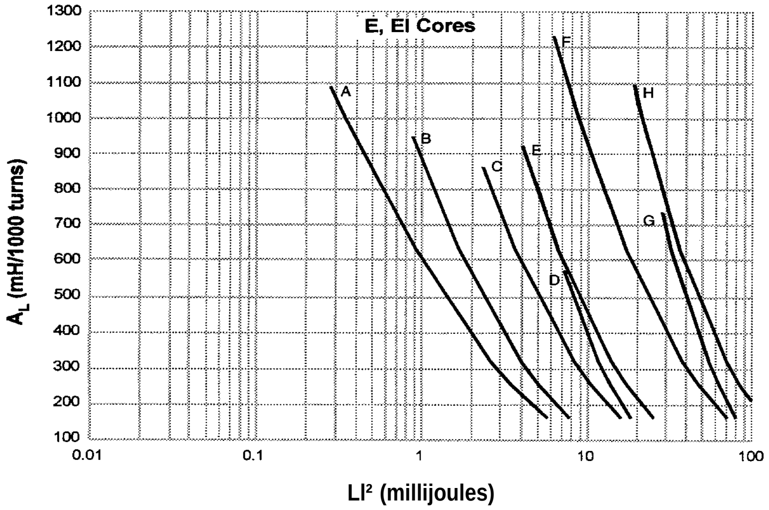
H- 42520 (EE)

J- 42810 (EE)
43013 (EE)

Ferrite DC Bias Core Selector Charts

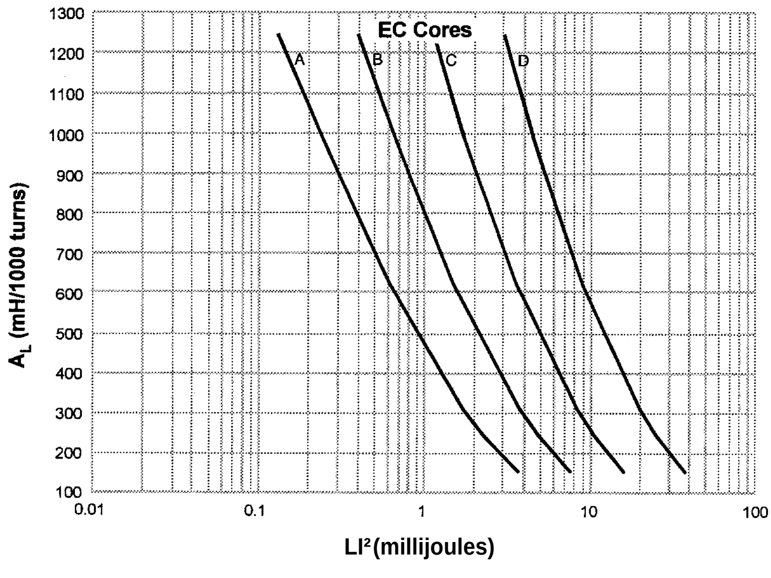


A- 42110 (EE)
B- 41709 (EE)
C- 41805 (EE, EI)
D- 42216 (EE, EI)
E- 44008 (EE, EI)
F- 43208 (EE, EI)
43618 (EE, EI)



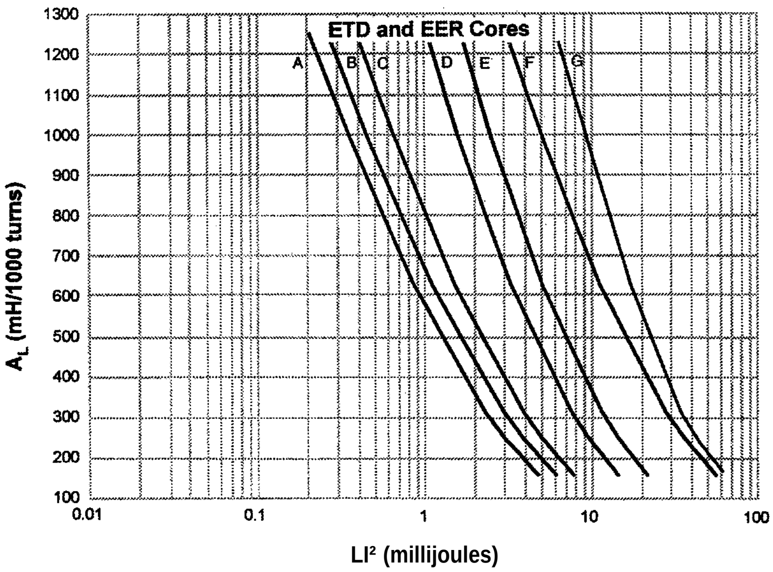
A- 44016 (EE)
B- 44011 (EE)
C- 44020 (EE)
D- 44308 (EE, EI)
E- 44022 (EE)
44924 (EE)
45021 (EE)
46016 (EE)

F- 45528 (EE)
45530 (EE)
47228 (EE)
48020 (EE)
G- 46410 (EE)
H- 49938 (EE, EI)

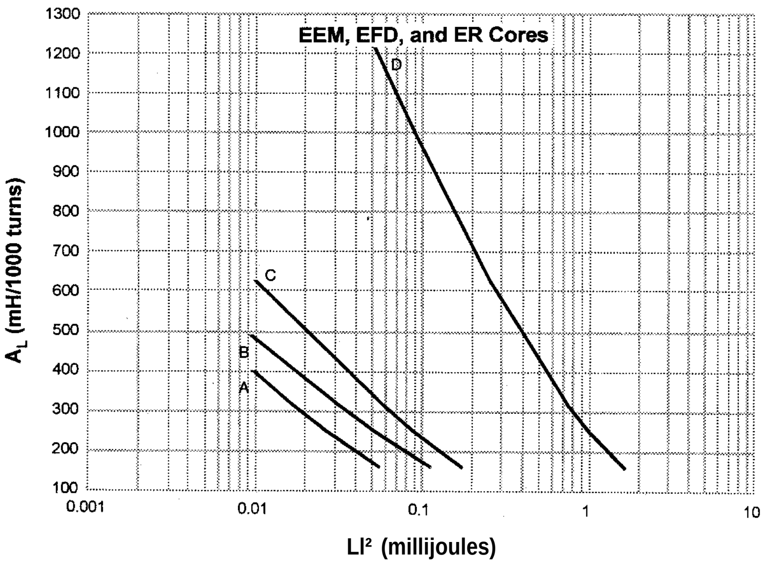


A- 43517
B- 44119
C- 45224
D- 47035

Ferrite DC Bias Core Selector Charts



A- 43434 (ETD34)
B- 43521 (EER35L)
C- 43939 (ETD39)
D- 44216 (EER42)
44444 (ETD44)
E- 44949 (ETD49)
F- 47054
G- 45959 (ETD59)



A- 41309 (EEM12.7)
40906

B- 42110
41515 (EFD15)

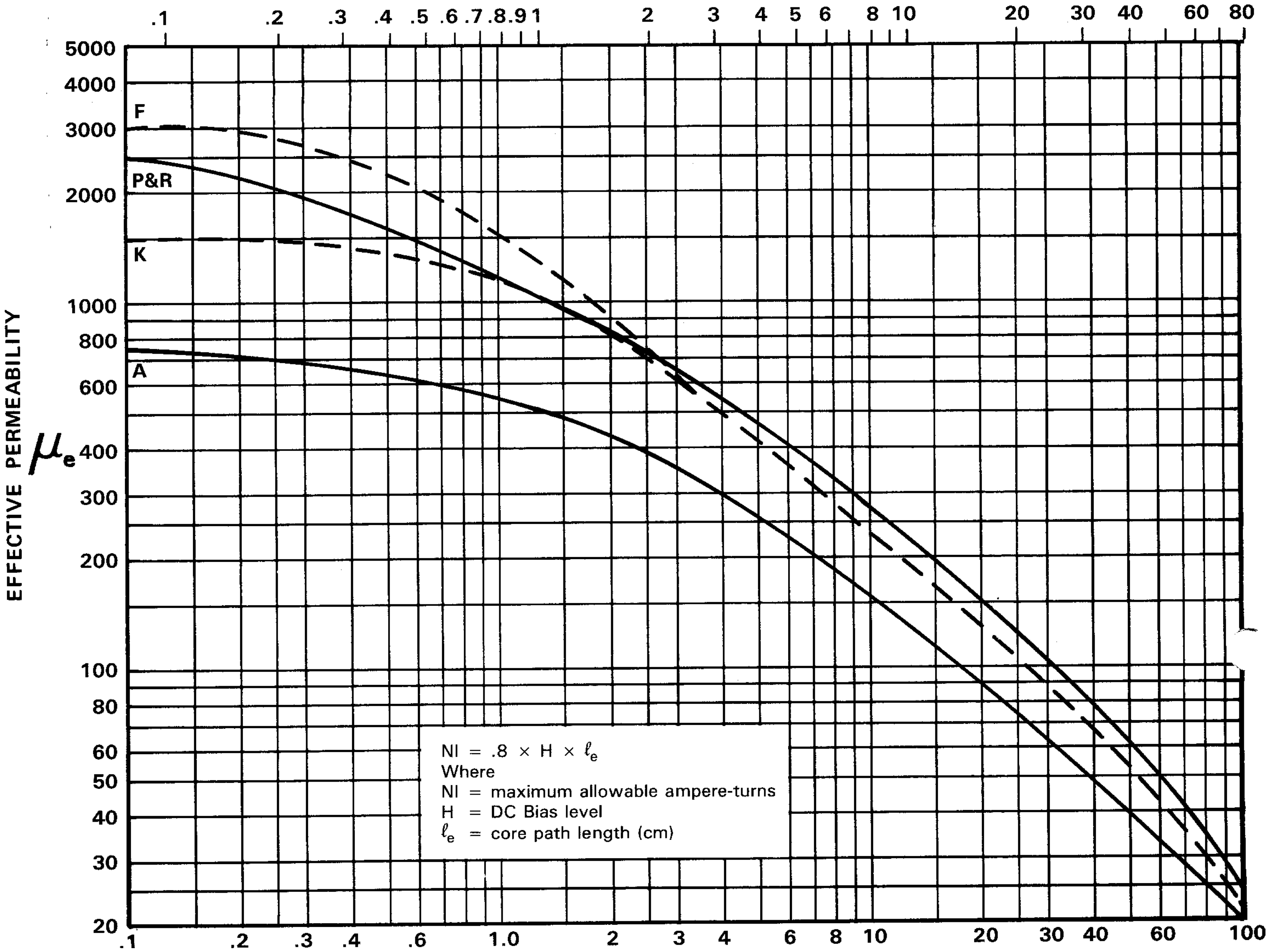
C- 41709

D- 42523 (EFD25)

Graph 4: For Gapped Applications — DC Bias Data

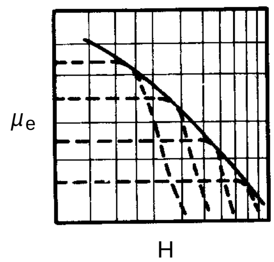
μ_e vs. H

H (Ampere-turns/cm)
(See bottom of graph for Oersteds)



H(OERSTED)
(See top of graph for Ampere-turns/cm)

The above curves represent the locus of points up to which *effective permeability* remains constant. They show the maximum allowable DC bias, in ampere-turns, without a reduction in inductance. Beyond this level, inductance drops rapidly.



Example: How many ampere-turns can be supported by an R-42213-A-315 pot core without a reduction in inductance value?

ℓ_e = 3.12 cm μ_e = 125
Maximum allowable H = 25 Oersted (from the graph above)
NI (maximum) = .8 x H x ℓ_e = 62.4 ampere-turns
OR (Using top scale, maximum allowable H = 20 A-T/cm.)
NI (maximum) = A-T/cm x ℓ_e
= 20 x 3.12
= 62.4 A-T

$$\mu_e = \frac{A_L \bullet \ell_e}{4\pi A_e}$$

$$\frac{1}{\mu_e} = \frac{1}{\mu_i} + \frac{\ell_g}{\ell_e}$$

A_e = effective cross sectional area (cm²)
A_L = inductance/1000 turns (mH)
μ_i = initial permeability
ℓ_g = gap length (cm)

Low Level Applications – Pot Cores

The information contained in this section is primarily concerned with the design of linear inductors for high frequency LC tuned circuits using Ferrite Pot Cores. Magnetics has arranged the data in this section for ease in (1) determining the optimum core for these LC circuits and (2) ordering the items necessary for any particular Pot Core assembly.

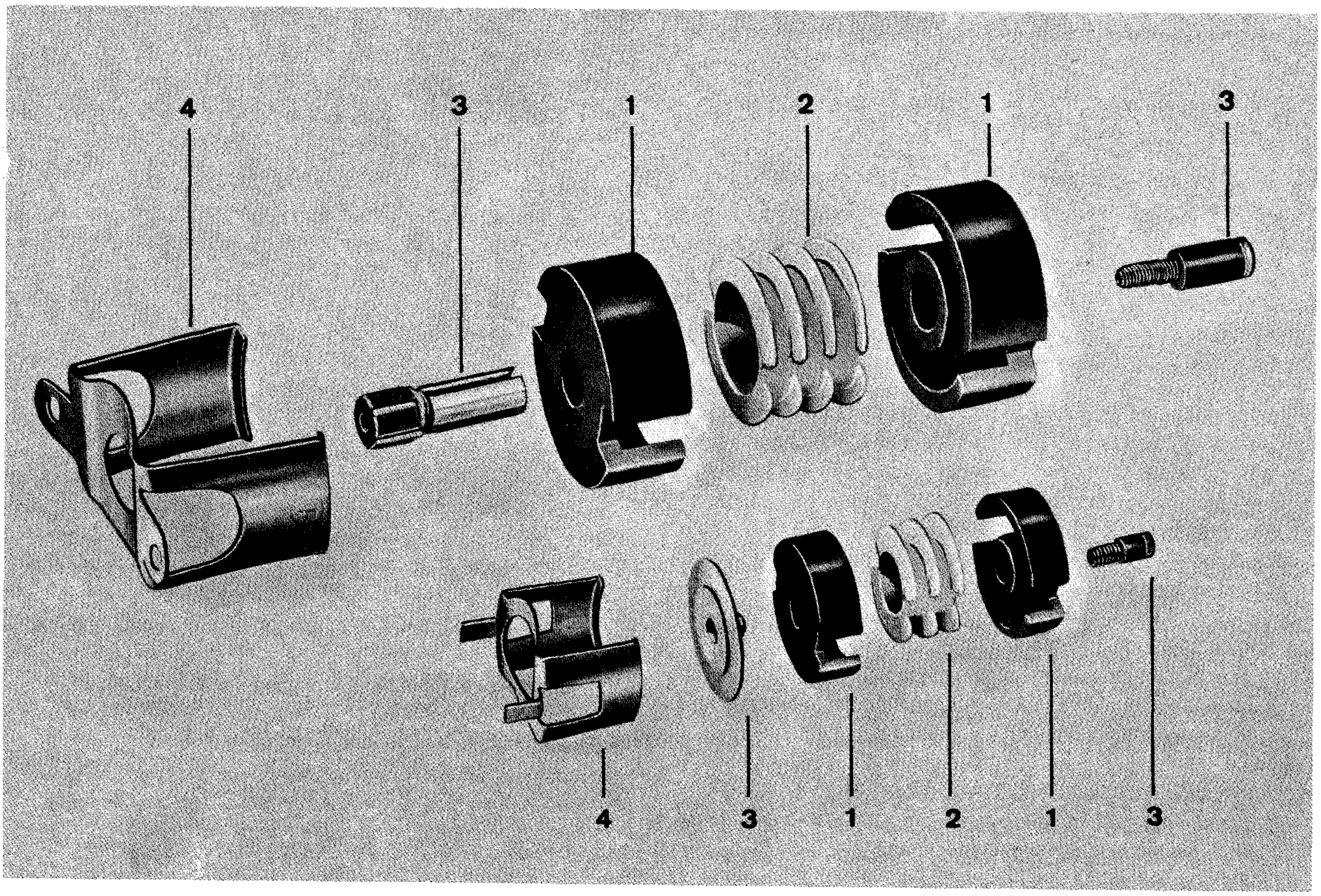
Featured are magnetic data, temperature characteristics, core dimensions, accessories, and other important design criteria. *Standard Q curves are available on special request, if needed.*

The data presented in this section are compiled mainly for selecting cores for high Q resonant LC circuits. However, much of this information can also be used to design pot cores into many other applications, including high frequency transformers, chokes, and other magnetic circuit elements.

Pot Core Assembly

A ferrite pot core assembly includes the following items:

(1) two matched pot core halves
(2) bobbin on which the coils are wound
(3) tuning assembly
(4) a clamp for holding the core halves together



The pot core shape provides a convenient means of adjusting the ferrite structure to meet the specific requirements of the inductor. Both high circuit Q and good temperature stability of inductance can be obtained with these cores. The self-shielded pot core isolates the winding from stray magnetic fields or effects from other surrounding circuit elements.

The effective permeability (μ_e) is adjusted by grinding a small air gap in the center post of the pot core. For transformers and some inductors, no ground air gap is introduced, and the effective permeability is maximized. The effective permeability of the pot core will always be less than the material initial permeability (μ_i) because of the small air gap at the mating surfaces of the pot core halves. For other inductors where stability of inductance, Q, and temperature coefficient must be closely specified, a controlled air gap is carefully ground into the center post of one or both of the pot core halves. When fitted together, the total air gap then will determine the effective permeability and control the magnetic characteristics of the pot core. Finer adjustment of the effective permeability (gapped pot core inductance) can be accomplished by moving a ferrite cylinder or rod into the air gap through a hol in the center post.

Magnetics ferrites are available in various initial permeabilities (μ_i) which for filter applications cover frequency ranges into the megahertz region. Magnetics produces a wide variety of pot core sizes which include fourteen (14) international standard sizes*. These range from 5 x 6 mm to 45 x 29 mm, these dimensions representing OD and height of a pair. Each pot core half is tested and matched with another half to produce a core with an inductance tolerance of ± 3% for most centerpost ground parts.

Advantages of Pot Core Assemblies

1. **SELF-SHIELDING**
   Because the wound coil is enclosed within the ferrite core, self-shielding prevents stray magnetic fields from entering or leaving the structure.

2. **COMPACTNESS**
   Self-shielding permits more compact arrangement of circuit components, especially on printed circuit boards.

3. **MECHANICAL CONVENIENCE**
   Ferrite pot cores are easy to assemble, mount, and wire to the circuit.

4. **LOW COST**
   As compared to other core materials, ferrites are easier to make in unusual configurations (such as pot cores), resulting in a lower cost component. In addition, winding a pot core is usually quick and inexpensive because coils can be pre-wound on bobbins. When other costs of assembly, mounting, wiring, and adjustment are added, the total cost is often less than with other core materials or shapes.

5. **ADJUSTABILITY**
   Final adjustment is accomplished by moving a threaded core in and out of the centerpost, and adjustment in the field is relatively easy as compared to any other type of construction.

6. **IMPROVED TEMPERATURE STABILITY AND Q**
   Air gaps inserted between the mating surfaces of the centerposts provide good temperature stability and high Q.

7. **WIDE CORE SELECTION**
   Many combinations of materials, physical sizes, and inductances offer the design engineer a large number of choices in core selection.

8. **LOW LOSSES AND LOW DISTORTION**
   Since ferrites have high resistivities, eddy current losses are extremely low over the applicable frequency range and can be neglected. Hysteresis losses can be kept low with proper selection of material, core size, and excitation level.

Special Advantages of Magnetics Pot Core Assemblies

1. **UNIQUE ONE PIECE CLAMP**
   Provides simple assembly of the two core halves. Easy bending action allows insertion of the core assembly into the clamp, and spring tension holds the assembly <u>rigidly</u> and <u>permanently</u> in place. Rivet, screw, or circuit board tab mounting is available.

2. **CHOICE OF LINEAR OR FLAT TEMPERATURE CHARACTERISTICS**
   Provides a close match to corresponding capacitors.

3. **CONSISTENCY AND UNIFORMITY**
   Modern equipment with closely controlled manufacturing processes produce ferrite pot cores that are magnetically uniform, not only within <u>one</u> <u>lot</u> but from <u>lot</u> to <u>lot</u>.

* IEC Publication No. 133 (1961).

Pot Core Design Notes Important Considerations

The selection of a pot core for use in LC resonant circuits and high frequency inductors requires a careful analysis of the design, including the following:

1. Operating frequency.
2. Inductance of the wound pot core assembly.
3. Temperature coefficient of the inductor.
4. Q of the inductor over the frequency range.
5. Dimensional limitations of the coil assembly.
6. Maximum current flowing through the coil.
7. Long term stability.

The important characteristics which strongly influence the above requirements are:

1. Relative loss factor — $\frac{1}{\mu iQ}$ This factor reflects the relative losses in the core and varies with different ferrite materials and changes in operating frequency. When selecting the proper material, it is best to choose the one giving the lowest $\frac{1}{\mu iQ}$ over the range of operating frequencies. In this way, the highest circuit Q can be expected. In a situation where the $\frac{1}{\mu iQ}$ curves may cross over or coincide at various frequencies, each ferrite material should be considered in view of all circuit parameters of importance, including size, temperature coefficient, and disaccommodation, as well as Q. With this analysis, little doubt is left concerning the optimum selection of a proper core material.

2. Inductance factor (A_L). The selection of this parameter is based on a logarithmic progressive series of values obtained by dividing a logarithmic decade into 5 equal parts (International Standardization Organization R5 series of preferred numbers). Since the AL values for the various core sizes are standard, they may be graphed or charted for ease of determining the required turns (N) to give the value of inductance needed. Graph 5 on page 5.8 is such a presentation and is derived from the formula $N = 10^3\sqrt{\frac{L}{A_L}}$ where L is the inductance in millihenries and A_L is the inductance factor. Pot cores with various A_L values are obtained by grinding closely-controlled air gaps in the centerposts of the cores. Small gaps are processed by gapping one core half. For larger gaps, both halves are gapped.

3. Temperature Coefficient (TC_e). The temperature coefficient of the pot core is important in LC tuned circuits and filters when attempting to stabilize the resonant frequency over a wide range of temperatures. This temperature coefficient (TC_e) is determined by the properties of the ferrite material and the amount of air gap introduced. Ferrite materials have been designed to produce gapped pot core temperature coefficients that balance the opposite temperature characteristics of polystyrene capacitors, or match similar flat temperature coefficients of silvered mica capacitors. Therefore, careful selection of both capacitors and pot cores with regard to temperature coefficient will insure the optimum temperature stability.

4. Quality Factor (Q)*. The quality factor is a measure of the effects of the various losses on circuit performance. From the designer's point of view, these losses should include core losses, copper losses, and winding capacitive losses. Therefore, Q will be affected greatly by the number and placement of the turns on the bobbin, and the type and size of wire used. At higher frequencies, litz wire would reduce the eddy current losses in the windings and produce a higher Q than solid wire. Q data include the effects of winding and capacitive losses, which, if removed, would produce significantly higher calculated Q values. Consequently, the Q curves represent more realistically the actual Q values that would be obtained from circuit designs.

5. Dimensional Limitations. Many circuit designs contain dimensional and weight limitations which restrict the size of the inductor and the mounting techniques used. Sometimes, minimum weight or volume is sacrificed to obtain better circuit performance.

6. Current Carrying Capacity. Inductive circuits containing ferrite pot cores are normally operated at extremely low levels of AC excitation to insure the best possible performance. However, the current flowing in the coil may be much higher than anticipated due to superimposed DC currents, or unexpected surges of AC. Therefore, the selection of the wire size used in an inductor design is influenced by both of these factors. Wire data is presented in this catalog as a guide in considering these operating conditions. Refer to Tables 5 and 6, page 5.9 or Graphs 6 and 7, pages 5.10 and 5.11.

6. Long Term Stability (DF_e). In critical inductive designs, especially resonant circuits, the designer must be concerned with long term drift in resonant frequency. This stability drift (or decrease in inductance), known as disaccommodation, can be calculated for each pot core size and inductance factor (A_L). It occurs at a logarithmic rate, and the long term change of inductance may be calculated from the formula:

$$\frac{\Delta L}{L} = DF_e \times \log\frac{t_2}{t_1}$$

where $\frac{\Delta L}{L}$ is the decrease in inductance between the times t_1 and t_2, DF_e is the Effective Disaccommodation Coefficient of the core selected, and t_1 is the elapsed time between manufacture of the core (stamped on shipping container) and its assembly into the circuit, while t_2 is the time from manufacture of the core to the end of the expected life of the device. Disaccommodation starts immediately after the core is manufactured as it cools through its Curie Temperature. At any later time as the core is demagnetized, or thermally or mechanically shocked, the inductance may increase to its original value and disaccommodation begins again. Therefore, consideration must be given to increases in inductance due to magnetic, thermal or physical shock, as well as decreases in inductance due to time. If no extreme conditioning is expected during the equipment life, changes in inductance will be small, because most of the change occurs during the first few months after manufacture of the core.

*Q curves referred to here are available on special request.

Limits on Excitation

Inductors designed using pot cores are usually identified as linear magnetic components because they are operated within the range of negligible change of effective permeability with excitation. To calculate suggested maximum AC excitation levels, use the following formula:

$$B = \frac{E_{rms} \times 10^8}{4.44\ A_e N f}$$

4.44 for sine wave
4.0 for square wave

where B = 200 gausses, the suggested conservative limit.
N = turns on pot core
f = operating frequency in hertz.
A_e = effective area of the pot core in cm².

Because superimposed DC current also affects linearity of inductance in pot cores, consideration for DC currents must also be given. The equation shown above must be modified to include effect of DC bias. The combined equation now becomes:

$$B_{(combined)} = \frac{E_{rms} \times 10^8}{4.44\ A_e N f} + \frac{N I_{dc} A_L}{10 A_e}$$

where B = 200 gausses, the suggested conservative limit.
I_{dc} = bias current in amperes.

See pages 4.15-4.19 for DC bias data on Magnetics power ferrites.

Material Selection

In most designs the ferrite material type can be obtained by knowing circuit operating frequencies. Refer to Graph 1 on page 2.3 which shows the $\frac{1}{\mu i Q}$ properties of Magnetics ferrites. The material type that gives the lowest $\frac{1}{\mu i Q}$ over the desired frequency range is normally selected.

Pot Core Selection

Various methods are used by engineers and designers to select the best pot core size for an application. These vary from a random trial approach to the more exact techniques involving the use of Q curves* or **ISO-Q contour curves*.** Included in this catalog are many graphs, charts, and tables that simplify selection techniques.

Q curves are developed from standard Q versus frequency measurements using coils with different numbers of turns and wire sizes. From the data obtained, standard Q curves are plotted and are available upon request. For convenience in selecting the optimum core, the TC_e and bobbin area are also listed with the Q graphs.

The only other remaining factors that must be determined are the exact wire size, dimensions, and the effects of disaccommodation. This information can be obtained from graphs, charts, or formulas referenced in this design section.

Depending on the requirements of the application, certain factors become more critical than others and will influence the design procedure used. For example, in an LC tuned circuit, if highest Q is required, larger cores must be used, and the C of the capacitor is usually obtained after determining the optimum L of the inductor. However, if size is critical, and maximum inductance is required to fit an available space, highest Q and optimum TC must be sacrificed. If all design factors are critical (as is now frequently demanded), each must be carefully evaluated with respect to its effect on the others to arrive at the best design.

The following is a review of several common types of application problems found in inductor design. Included is a suggested procedure to be followed to obtain an optimum solution:

Application 1. Frequency and Inductance are specified. Smallest size is important. Q or temperature stability is not critical.

Solution. For the frequency range specified, select the best suited core material, using the $\frac{1}{\mu i Q}$ curves.

Find the smallest core size with the highest A_L available in the material selected, by referring to the data listed for each size. For this A_L, determine the turns required to meet the inductance specified, using Graph 5, page 5.8 or equation $N = 10^3\sqrt{L/A_L}$.

Knowing the core size, select the maximum wire size from Graph 6, page 5.10. As an alternate procedure, divide the turns by the Bobbin area to obtain the turns per square inch and refer to the Wire Tables page 5.9 for maximum wire size. If this wire size doesn't exceed the one that can economically be used, the design is complete. If it does, select the next larger core size and the highest AL, and recheck until a satisfactory wire size is obtained.

Application 2. Frequency is specified. Q must be maximum. The L consistent with this maximum Q will be used to determine the C in a resonant circuit. Size or temperature stability is not critical.

*Available upon request.

Solution. Select the core material best suited for the frequency range using the $\frac{1}{\mu_i Q}$ curves. Scan the Q curves for the material chosen to find the core size and A_L that give the highest Q over the frequency range specified. Record the core size, A_L, and inductance L shown for that Q.

For the A_L chosen, obtain the turns required to produce the inductance L from Graph 5, page 5.8 or use equation $N = 10^3\sqrt{L/A_L}$

Knowing the core size, select the wire size from Graph 7, page 5.11. The use of litz wire will usually increase Q.

Further increases in Q may be accomplished with the use of double or triple section bobbins or by spacing the winding away from the air gap.

Determine the C required for resonance from $f = \frac{1}{2\pi\sqrt{LC}}$

Application 3. Frequency, inductance, minimum Q, temperature coefficient, and maximum size are specified.

Solution. Using the $\frac{1}{\mu_i Q}$ curves, select the core material best suited for the frequency range. Scan the Q curves for the material chosen and select the core sizes and A_L's that give the desired TC_e. List these cores.

Using the Q curves, eliminate those sizes that do not meet the minimum Q at the frequency and inductance value specified. From the remaining cores, choose the smallest size. Determine the number of turns required to meet the inductance by using Graph 5, page 5.8 or use equation $N = 10^3\sqrt{L/A_L}$. As an alternate procedure, divide the turns by the bobbin area to obtain the turns per square inch, and refer to the Wire Tables page 5.9 for the maximum wire size. The use of litz wire will usually increase Q.

Check the design dimensions against the core size. If the maximum size is exceeded, Q and inductances must be re-evaluated, and a new core size selected.

Note: While not mentioned in the above applications, critical LC tuned circuits must remain stable over long periods of time. Disaccommodation, the decrease of inductance over time, can be calculated from the formula, $\frac{\Delta L}{L} = DF_e \log\frac{t_2}{t_1}$, as previously noted. From the calculated decrease of inductance, the shift in the resonant frequency can be predicted. Disaccommodation can be ignored if the shift is within limits.

Core Selection Chart

These tables summarize the design procedures discussed in the preceding section.

Application 1. Frequency and Inductance Specified.
Size must be minimum.

Step	Operation	Reference
1.	Select core material	Graph 1, page 2.4
2.	List smallest core size with highest A_L	See core size pages
3.	Find turns required	Graph 5, page 5.8 $N = 10^3\sqrt{L/A_L}$
4.	Select wire size	Tables 5 & 6, page 5.9 Graph 6, page 5.10 $N = 10^3\sqrt{L/A_L}$
5.	If too small to wind economically, repeat steps 2, 3, and 4 for next larger core size.	

Application 2. Frequency specified with Q to be maximum.
C to be determined from L when Q is maximum.

Step	Operation	Reference
1.	Select core material	Graph 1, page 2.3
2.	Find core with highest Q over frequency range	Q curves*
3.	Record core size, A_L, and inductance L	See core size pages
4.	Obtain turns required	Graph 5, page 5.8 $N = 10^3\sqrt{L/A_L}$
5.	Select wire size	Graph 6, page 5.10 Tables 5 & 6, page 5.9
6.	Determine C for resonance	Formula $C = \frac{1}{(2\pi f)^2 L}$

*Available upon request.

Application 3. All requirements equally important.

Steps	Operation	Reference
1.	Select core material	Graph 1, page 2.3
2.	List all cores with acceptable TCe values	See core size pages
3.	From cores in step 2 select those with acceptable Q at the inductance and frequency specified.	Q curves*
4.	Using the smallest core selected in 3, find the turns N	Graph 5, page 5.8 $N = 10^3\sqrt{L/A_L}$
5.	Select wire size	Tables 5 and 6, page 5.9 Graph 6, page 5.10
6.	Check size of core against design dimensions	See core size pages
7.	If size is exceeded, re-evaluate Q and L and repeat steps 2 to 6 to select a smaller core	
8.	Calculate decrease of inductance for life of circuit	Q curves or Paragraph 7, page 5.3
9.	Evaluate AC and DC bias currents to insure wire size is adequate.	Wire Tables 5 and 6, page 5.9 Graphs 6 and 7, pages 5.10 and 5.11

EXAMPLE:

Design a 5 mh inductor for use at 100 KHz having Q of 400 minimum. The inductor must fit into a space 1” x 1” x ¾” and shall have a temperature coefficient TCe of 220 nominal.

1. Select core material - “D” rather than “G” because of the temperature coefficient requirement.
2. List all cores with acceptable TCe values.

D-40704-160A $_L$	D-41 408-250A $_L$
D-40905-160A $_L$	D-41811-315A $_L$
D-41107-250A $_L$	D-42213-400A $_L$

3. Select those with acceptable Q at the inductance and frequency.
   D-41811-315A $_L$
4. Using the smallest core from step 3, find the turns N.
   N = 130 turns
5. Select wire size. A single section bobbin for the 41811 core has an area of .026 sq. in. Dividing turns (130) by area (.026), turns per square inch = 5000. From the wire table for litz wire, the largest wire size that can be used is 20/44.
6. Check dimensions - 41811 size core has a .709” diameter and .416” height which meet the dimensional requirements.

Pot Core Assembly Notes

Magnetics ferrite pot cores can be assembled with or without clamping hardware or tuning devices.

Mounting clamps are available for the 40905, 41107, 41408, 41811, 42213, 42616, 43019, 43622, and 44229 pot core sizes. These clamps normally eliminate the need to cement the pot core halves together. The mating surfaces of the pot core must be cleaned to remove moisture, grease, dust, or other foreign particles, before clamping or cementing.

If the cementing method is chosen, a small amount of cement is placed on the mating surface of the pot core skirt, being careful to keep the centerpost free of all cement. The pot core halves are brought together and rotated together under slight pressure to distribute the cement evenly around the skirt. The halves are separated and the wound bobbin is set in place. A small amount of cement is now placed on the exposed flange of the bobbin to bond it in the pot core assembly and thus insure no movement. The other core half is replaced, the centerpost holes and wire apertures aligned, and the unit clamped together in a pressure jig. Permanent bonding is accomplished by curing the cement at elevated temperatures according to the manufacturer’s recommendations. After curing, storage for a minimum of 24 hours, and heat cycling between room temperature and 70°C may be required before final testing or tuning is completed.

The tuning adjusters can be inserted into the pot core immediately after the cemented core halves have been cured and the assembly can then be heat cycled. Some adjusters require insertior of the base into the centerpost hole prior to assembly of the pot core into the clamp when a clamp is used for mounting. The ad-

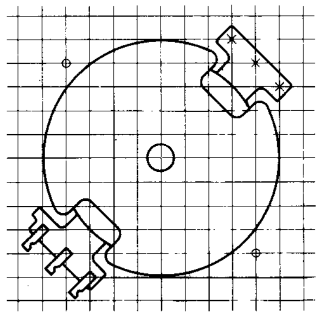
*available on request

juster is usually made in two parts - the plastic base with a threaded hole, and a ferrite cylinder imbedded in a plastic screw. The base is pressed into the centerpost of the pot core, and the plastic screw is turned into the base until the ferrite cylinder nters the air gap. Tuning is completed when the inductance of the pot core assembly reaches the proper value. If this initial adjustment is expected to be the final one, cementing is recommended to prevent accidental detuning. If precise inductance values are expected, final tuning should not be completed earlier than 24 hours after the pot core assembly has been cured or clamped.

"TB-P" bases, which are polypropylene, may be etched in order to roughen the adhering surface and improve the bonding that is achieved.

Plastic screw drivers are available upon request for use in final tuning.

FIGURE 1



PRINTED CIRCUIT BOBBINS AND MOUNTING HARDWARE

Many sizes in the standard pot cores can be supplied with printed circuit board bobbins. The grid pattern (Figure 1) illustrates the location of 6 pin type bobbins. The soldering pins are arranged to fit a grid of 0.1", and they will also fit printed circuit boards with 2.50 mm grids. The pin length is sufficient for a board thickness up to .187". Terminal pin details are illustrated in Figure 2. The board holes should be .046" + .003" in diameter (#56 drill). The bobbin should be cemented the lower pot core half.

For some core types, printed circuit board mounting clamps are also available. A cross section of a completed core assembly using clamps is shown in Figure 3. When clamps are not available, the pot core halves must be cemented together.

Printed circuit board hardware for EP, RM and RS cores is described in the sections covering these core types.

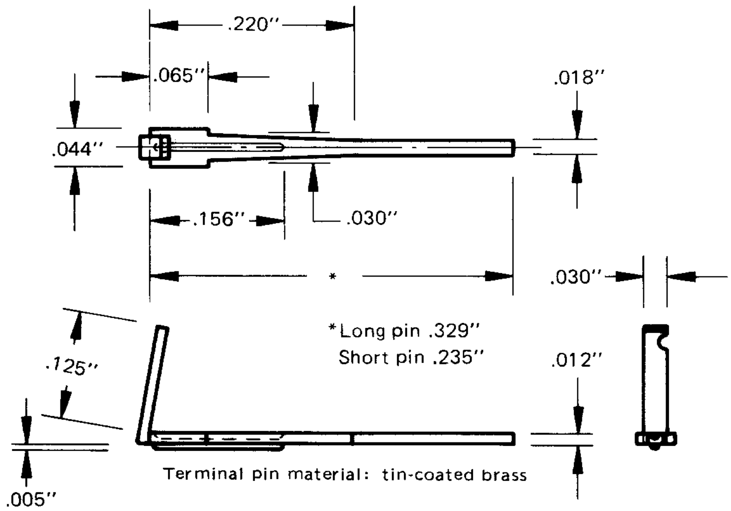


FIGURE 2

PRINTED CIRCUIT BOBBINS SOLDERING INSTRUCTIONS

1. A solder pot should be used to solder the leads to the terminals. Preferred solder is 63/37 tin/lead eutectic. The solder temperature should be between 275°-300°C. Lower or higher temperatures will both damage the bobbin. Modern soldering techniques commonly use temperatures *in excess of the softening points of all thermoplastic bobbin materials.* Extreme care is required to prevent loosening of the terminals during soldering.
2. Insulation should be removed from the ends of the wire before soldering. This is especially important when litz wire is used. The preferred method is by burning.
3. Dip wound terminals into liquid soldering flux. A rosin based flux in alcohol solution should be used. Allow flux to air dry.
4. The bobbin should be immersed only far enough to cover the terminals.
5. The part should be immersed in the solder for 2-4 seconds, depending on the size of the wire used.

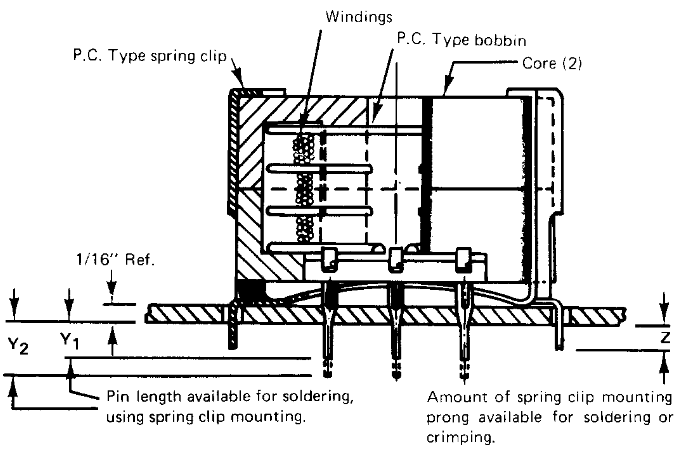
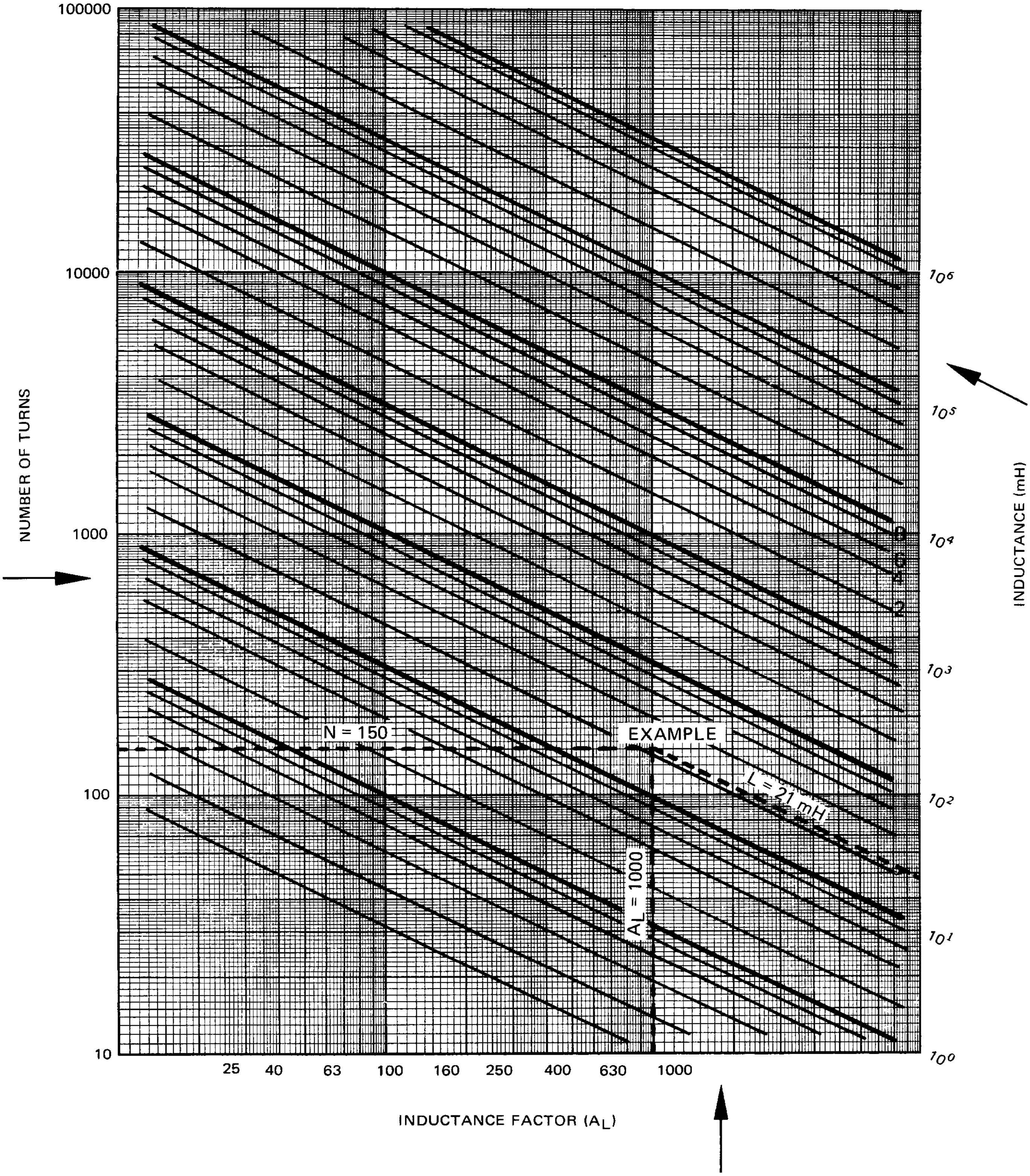
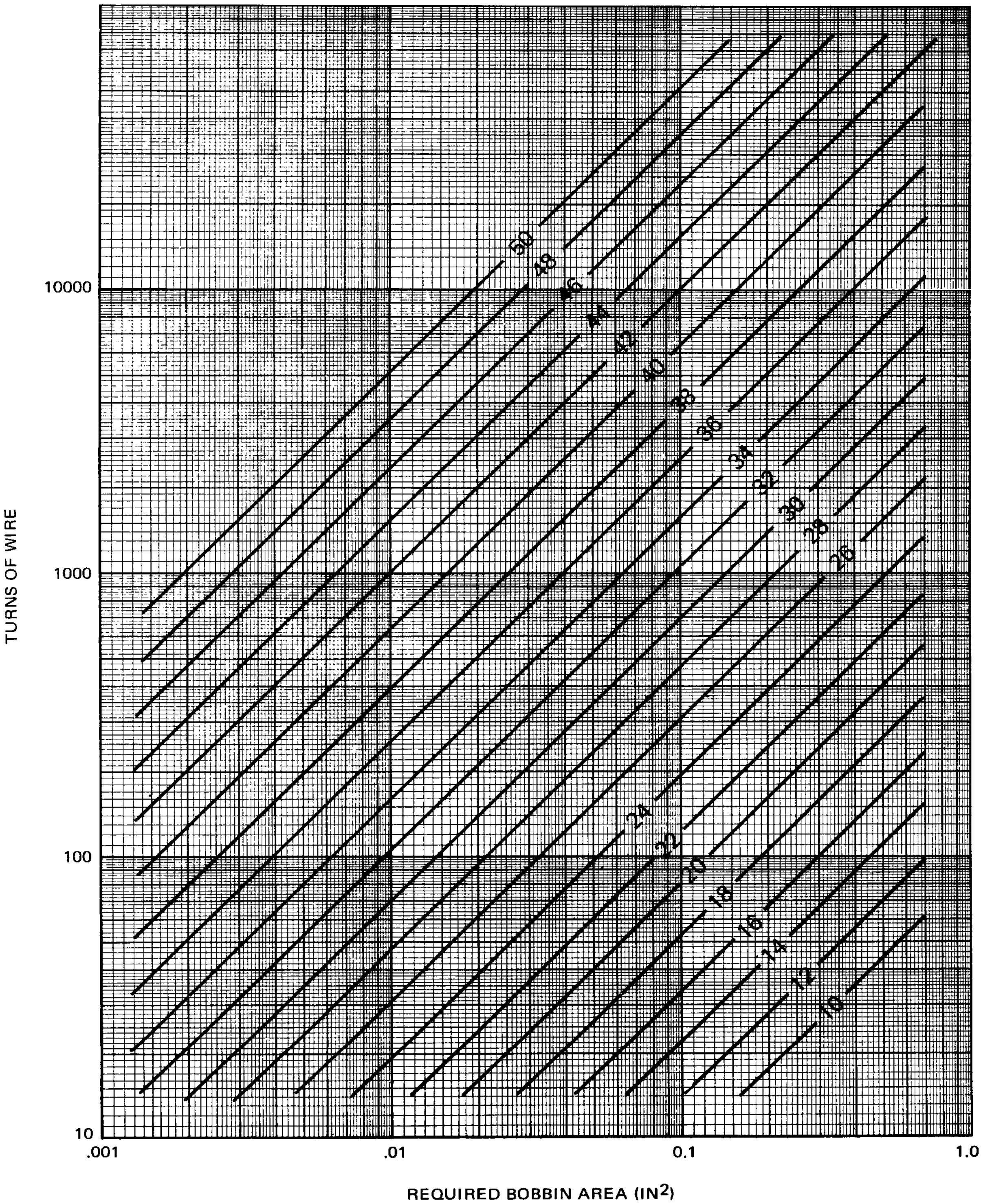


FIGURE 3

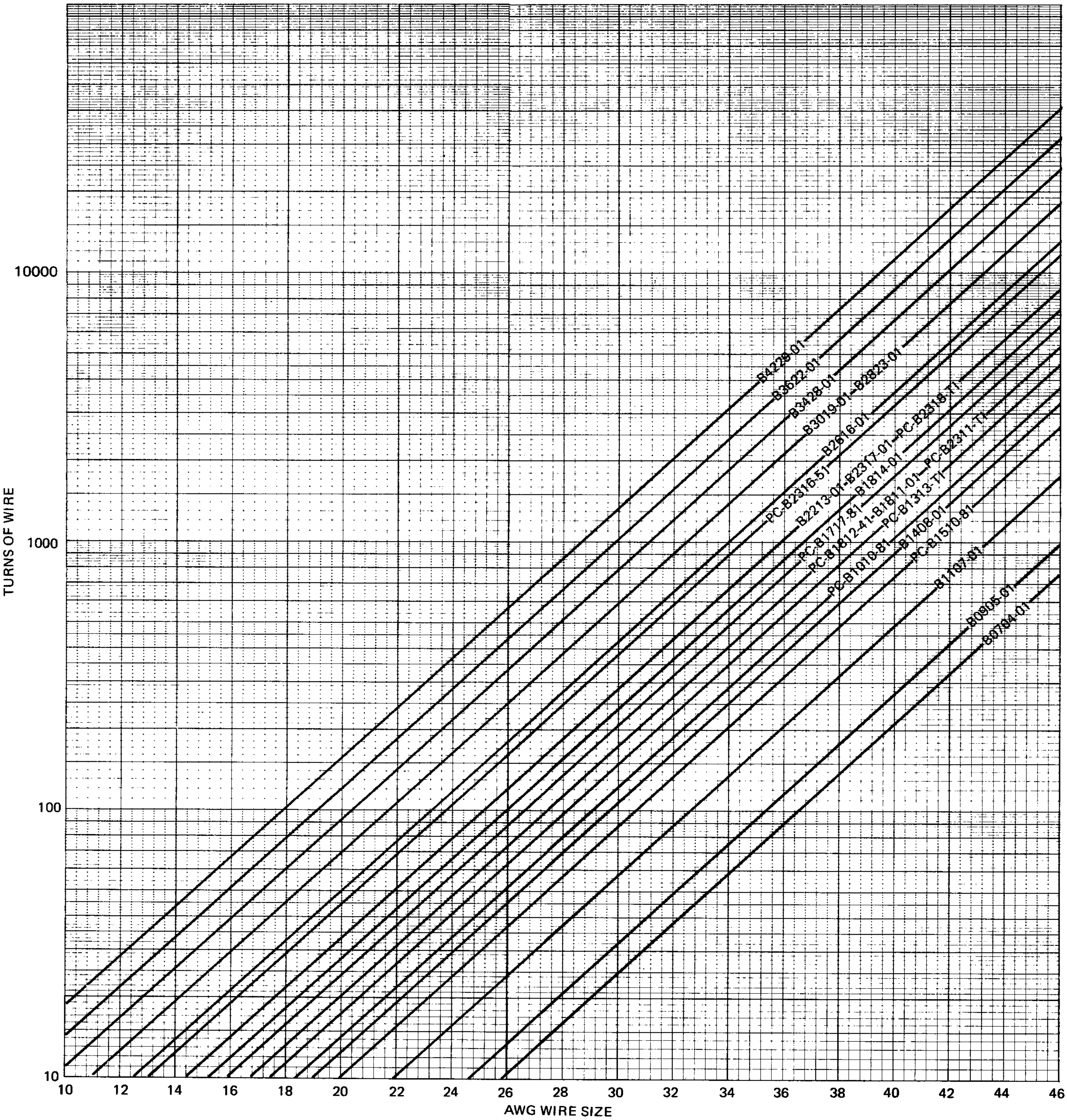
Graph 5 — Turns/Inductance Factor/Inductance Graph



Graph 6 – Bobbin Area/Turns/AWG Wire Size Graph (all core geometries)



Graph 7 – Wire Size/Maximum Turns/Bobbin Size Graph (pot cores)



To obtain turns on two section bobbins, multiply answer from graph by .47 and for three section bobbins multiply by .30. Exception: multiply turns obtained from graph by .42 for the B1107-02 and .44 for the B1408-02 bobbin.

Table 5 — Magnet Wire

Wire Size	Wire Area (Max.)* Heavy		Turns**		Resistance	Current Capacity (ma)	
AWG	Circular Mils	$cm^2\ 10^{-3}$	per in^2	per cm^2	Ohms/1000'	@ 750 Cir. Mil/amp	@ 500 Cir. Mil/amp
10	11,470	58.13	89	13.8	.9987	13,840	20,768
11	9,158	46.42	112	17.4	1.261	10,968	16,452
12	7,310	37.05	140	21.7	1.588	8,705	13,058
13	5,852	29.66	176	27.3	2.001	6,912	10,368
14	4,679	23.72	220	34.1	2.524	5,479	8,220
15	3,758	19.05	260	40.3	3.181	4,347	6,520
16	3,003	15.22	330	51.2	4.020	3,441	5,160
17	2,421	12.27	410	63.6	5.054	2,736	4,100
18	1,936	9.812	510	79.1	6.386	2,165	3,250
19	1,560	7.907	635	98.4	8.046	1,719	2,580
20	1,246	6.315	800	124	10.13	1,365	2,050
21	1,005	5.094	1,000	155	12.77	1,083	1,630
22	807	4.090	1,200	186	16.20	853	1,280
23	650	3.294	1,500	232	20.30	681	1,020
24	524	2.656	1,900	294	25.67	539	808
25	424	2.149	2,400	372	32.37	427	641
26	342	1.733	3,000	465	41.0	338	506
27	272	1.379	3,600	558	51.4	259	403
28	219	1.110	4,700	728	65.3	212	318
29	180	0.9123	5,600	868	81.2	171	255
30	144	0.7298	7,000	1,085	104	133	200
31	117	0.5930	8,500	1,317	131	106	158
32	96.0	0.4866	10,500	1,628	162	85	128
33	77.4	0.3923	13,000	2,015	206	67	101
34	60.8	0.3082	16,000	2,480	261	53	79
35	49.0	0.2484	20,000	3,100	331	42	63
36	39.7	0.2012	25,000	3,876	415	33	50
37	32.5	0.1647	32,000	4,961	512	27	41
38	26.0	0.1318	37,000	5,736	648	21	32
39	20.2	0.1024	50,000	7,752	847	16	25
40	16.0	0.0811	65,000	10,077	1,080	13	19
41	13.0	0.0659	80,000	12,403	1,320	11	16
42	10.2	0.0517	100,000	15,504	1,660	8.5	13
43	8.40	0.0426	125,000	19,380	2,140	6.5	10
44	7.30	0.037	150,000	23,256	2,590	5.5	8
45	5.30	0.0269	185.000	28,682	3,348	4.1	6.2

Table 6 — Litz Wire

Litz Wire Size	Turns***	
	per in^2	per cm^2
5/44	28,000	4,341
6/44	25,000	3,876
7/44	22,000	3,410
12/44	13,000	2,016
20/44	7,400	1,147
30/44	4,000	620
40/44	3,000	465
50/44	2,300	356
60/44	1,900	294

Litz Wire Size	Turns***	
	per in^2	per cm^2
72/44	1,500	232
80/44	1,400	217
90/44	1,200	186
100/44	1,100	170
120/44	900	140
150/44	700	108
180/44	500	77
360/44	250	38

*Areas are for maximum wire area plus maximum insulation buildup.

**Based on a typical machine layer wound coil.

***Based on a typical layer wound coil.

Symbols and Definitions

Symbol	Units	Definition
μ	- -	Permeability—The ratio of magnetic flux density in gausses to magnetic field strength in oersteds. $$\mu = \frac{B}{H}$$
μ_i	- -	Initial Permeability—The value of the permeability at very low magnetic field strengths. $$\mu_i = \lim_{H \to O} \frac{B}{H}$$
μ_e	- -	Effective Permeability—If a magnetic circuit is not homogeneous (i.e., contains an air gap), the effective permeability is the permeability of a hypothetical homogeneous (ungapped) structure of the same shape, dimensions, and reluctance, that would give the inductance equivalent to the gapped structure.
A_L	millihenries per 1000 turns or nanohenries/turn²	Inductance factor—In a wound core, the inductance per unit turn when L is in henries. More often, when L is expressed in millihenries, A_L is the inductance as measured using a thousand turn coil. When calculating for other turns, use: L (mH) = $A_L n^2/1000^2$.
TC	/°C	Temperature Coefficient—The relative change in permeability per °C when measured at two different temperatures. $$TC = \frac{\mu_2 - \mu_1}{\mu_1 (T_2 - T_1)}$$
TF	/°C	Temperature Factor—The temperature coefficient of a material per unit of permeability. $$TF = \frac{TC}{\mu_i}$$
TC_e	/°C	Effective Temperature Coefficient—The actual temperature coefficient of a magnetic structure whose material permeability has been reduced to μe by gapping. $$TCe = TF \times \mu e$$
DA	- -	Disaccommodation—The relative decrease in permeability of a magnetic material with time after magnetic conditioning (demagnetization). $$DA = \frac{\Delta\mu}{\mu_i} \Big/ \log\frac{t_2}{t_1}$$ t_1 = time from demagnetization to 1st measurement t_2 = time from demagnetization to 2nd measurement For each decade of time, when $t_2 = 10t_1$ $$DA = \frac{\Delta\mu}{\mu_i}$$
DF	- -	Disaccommodation Factor—The disaccommodation of a material per unit of permeability. $$DF = \frac{DA}{\mu_i}$$

Symbol	Units	Definition
DF_e	- -	Effective Disaccommodation Coefficient—The actual disaccommodation of a magnetic circuit whose material permeability has been reduced to μe by gapping. $DF_e = DF \times \mu_e$
Q	- -	Q Factor—The efficiency of an inductor, that is the ratio of series inductive reactance to loss resistance. $Q = \frac{\omega L_s}{R_s}$
tan δ	- -	Loss angle—Deviation from ideal phase angle (90°) due to losses. $\tan\delta = \frac{R_s}{\omega L_s} = \frac{1}{Q}$
$\frac{\tan\delta}{\mu_i}$	- -	Relative loss factor—Losses per unit of permeability. Figure of merit of a material. $\frac{\tan\delta}{\mu_i} = \frac{1}{\mu_i Q}$
Ch	/gausses	Hysteresis coefficient—The coefficient in the Legg* * equation which separates the hysteresis losses from the eddy current and residual losses. $\frac{R_s}{\mu_i f L_s} = C_h B + C_e f + C_r$ This coefficient can be evaluated by noting the variation of series resistance with B.
$\frac{C_h}{\mu_i^2}$	/gausses	Relative Hysteresis Factor. The hysteresis coefficient normalized to unit permeability so that it is strictly a material property.
$C_{h(e)}$	/gausses	Effective Hysteresis Coefficient—The actual hysteresis loss in a magnetic structure whose permeability has been reduced to μe by gapping. $C_{h(e)} = \frac{C_h}{\mu_i^2} \times \mu_e^2$
B	gausses	Flux density—The magnetic flux in maxwells per cm² of cross sectional area.
B_{max}	gausses	The flux density at high field strengths (normally 25 oersteds).
	Teslas	10^4 gausses = 1 Tesla
H	oersteds	Field strength—The externally applied magnetizing field in oersteds.
H_c	oersteds	Coercive force—The reverse magnetic field needed to reduce a magnetically saturated structure from remanence to zero magnetic induction.
	amp-turns/m	1 oersted = 79.5 amp-turns/m
L	henries	Inductance—The magnetic flux linkages in maxwells-turns per ampere of magnetizing current. $L = -N \frac{d\Phi}{di}$
l_e	cm	Effective magnetic path length—In a structure containing a non-uniform cross section, the effective magnetic path length is that length of a similar structure with uniform cross section which is equivalent to the first for purposes of magnetic calculations.
A_e	cm²	Effective cross-sectional area—In a structure containing a non-uniform cross section, the effective magnetic cross section is the area of a structure with uniform cross section which is equivalent to the first for purposes of magnetic calculations.
V_e	cm³	Effective magnetic volume.
T_c	° c	Curie Temperature—Temperature at which a ferromagnetic material loses its ferromagnetism and becomes paramagnetic (μ approaches 1).

****V.E. Legg, Magnetic Measurements at Low Flux Densities Using the Alternating Current Bridge, Bell System Technical Journal, 15, 39 (1936)**