Section 4

Power Design

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.

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.

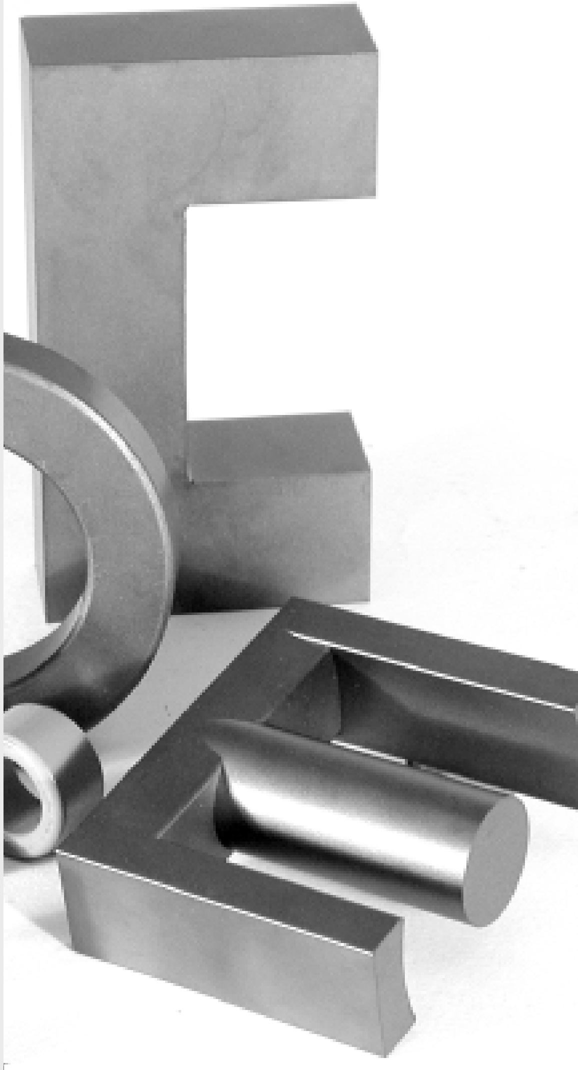
Ferrite cores may also be used in fly-back transformer designs, which offer low core cost, low circuit cost and high voltage capability. Powder cores (MPP, High Flux, Kool Mµ®) offer soft saturation, higher B_{max} and better temperature stability and may be the best choice in some flyback applications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



Materials and Geometries

CORE MATERIALS

F, P, 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.

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

FERRITE
POWER MATERIALS SUMMARY

		F	P	R	J	W+
μ_i (20 gauss)	25°C	3,000	2,500	2,300	5,000	10,000
μ_p (2000 gauss)	100°C	4,600	6,500	6,500	5,500	12,000
Saturation Flux Density (B_m Gauss)	25°C	4,900	5,000	5,000	4,300	4,300
	100°C	3,700	3,900	3,700	2,500	2,500
Core Loss (mw/cm^3) (Typical) @100 kHz, 1000 Gauss	25°C	100	125	140		
	60°C	180	80*	100		
	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.

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.

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.

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.

PLANAR E CORES

Planar E cores are offered in all of the IEC standard sizes, as well as a number of other sizes. Magnetics R material is perfectly suited to planar designs due to its low AC core losses and minimum losses at 100°C. Planar designs typically have low turns counts and favorable thermal dissipation compared with conventional ferrite transformers, and as a consequence the optimum designs for space and efficiency result in higher flux densities. In those designs, the performance advantage of R material is especially significant.

The leg length and window height (B and D dimensions) are adjustable for specific applications without new tooling. This permits the designer to adjust the final core specification to exactly accommodate the planar conductor stack height, with no wasted space. Clips and clip slots are avail-

Materials and Geometries

able in many cases, which is especially useful for prototyping. I-cores are also offered standard, permitting further flexibility in design. E-I planar combinations are useful to allow practical face bonding in high volume assembly, and for making gapped inductor cores where fringing losses must be carefully considered due to the planar construction.

EC, ETD, EER AND ER 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.

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 CORES	DOUBLE SLAB, RM CORES	EP CORES	PQ CORES	E CORES	EC, ETD, EER, ER CORES	TOROIDS
See Catalog Section	6	7-8	9	10	11	12	13
Core Cost	High	High	Medium	High	Low	Medium	Very Low
Bobbin Cost	Low	Low	High	High	Low	Medium	None
Winding Cost	Low	Low	Low	Low	Low	Low	High
Winding Flexibility	Good	Good	Good	Good	Excellent	Excellent	Fair
Assembly	Simple	Simple	Simple	Simple	Simple	Medium	None
Mounting Flexibility**	Good	Good	Good	Fair	Good	Fair	Poor
Heat Dissipation	Poor	Good	Poor	Good	Excellent	Good	Good
Shielding	Excellent	Good	Excellent	Fair	Poor	Poor	Good

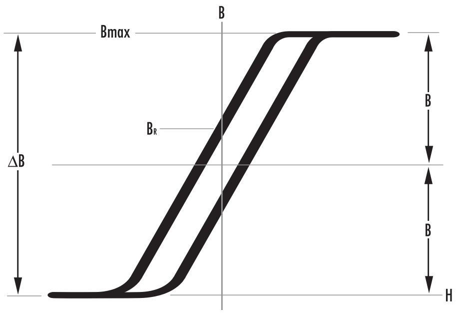
** Hardware is required for clamping core halves together and mounting assembled core on a circuit board or chassis.

General Formulas

TRANSFORMER CORE SIZE SELECTION

The power handling capacity on 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:

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

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

Where:
E=applied voltage (rms)
B=flux density in gauss
Ac=core area in cm^2
N=number of turns
f=frequency in Hz
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
K=winding factor
I=current (rms)
P_i=input power
P_o=output power
e=transformer efficiency

Solving (1) for NAc

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

The winding factor

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

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

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

In addition:

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

Thus:

$E\,Aw = EIC = P_i\,C = P_o\,C/e$

Substituting for EAw in (4), we obtain:

$$WaAc = \frac{P_oC \times 10^8}{4eB\,fK}$$

Assuming the following operational conditions:

$C = 4.05 \times 10^{-3} cm^2$/Amp (square wave) and
  $2.53 \times 10^{-3} cm^2$/Amp (sine wave) for toroids
$C = 5.07 \times 10^{-3} cm^2$/Amp (square wave) and
  $3.55 \times 10^{-3} cm^2$/Amp (sine wave) for pot cores and
  E-U-I cores.
e= 90% for transformers
e= 80% for inverters (including circuit losses)
K= 0.30 for pot cores and E-U-I cores (primary side only)
K= 0.20 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 pgs 5.8.

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

$$WaAc = \frac{k' P_0 \times 10^8}{Bf}, \text{ Where } k' = \frac{C}{4eK}$$

For square wave operation

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

A core selection chart (Table 3) using WaAc can be found on page 4.7. In addition a A core selection procedure which varies by topology can also be found on page 4.8. This procedure is based on the book "Switching Power Supply Design" by A.I. Pressman. While the formula above allows WaAc to be adjusted based on selected core geometry, the Pressman approach uses topology as the key consideration and allows the designer to specify current density.

GENERAL INFORMATION

An ideal transformer is one that offers minimum core loss while requiring the least amount of space. The core loss of a given core is directly effected by the flux density and the frequency. Frequency is the most important characteristic concerning a transformer. Faraday's Law illustrates that as frequency increases, the flux density decreases proportionately. Core losses decrease more when the flux density drops than when frequency rises.

For example, if a transformer were run at 250 kHz and 2 kG on R material at 100°C, the core losses would be approximately 400 mW/cm^3. If the frequency were doubled and all other parameters untouched, by virtue of Faraday's law, the flux density would become 1kG and the resulting core losses would be approximately 300mW/cm^3.

Typical ferrite power transformers are core loss limited in the range of 50-200mW/cm^3. Planar designs can be run more aggressively, up to 600 mW/cm^3, due to better power dissipation and less copper in the windings.

Specific Circuit Examples

CIRCUIT TYPES

Some general comments on the different circuits are:

The push-pull circuit is efficient because it makes bi-directional 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

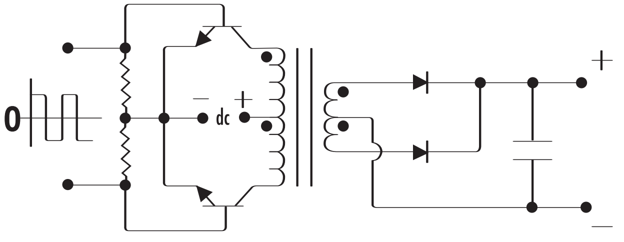
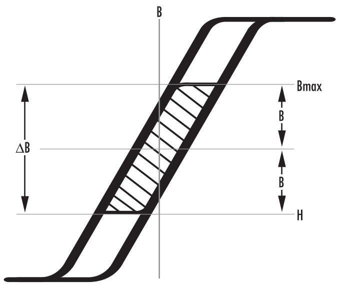


FIGURE 2B – HYSTERESIS LOOP OF MAGNETIC

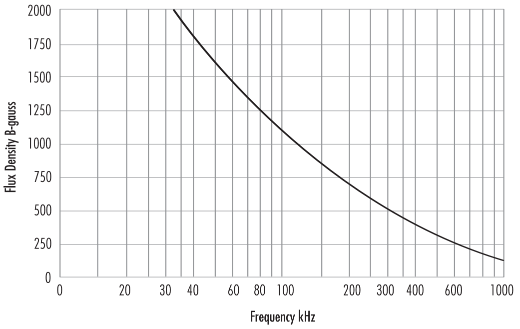


CORE IN PUSH-PULL CIRCUIT

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.

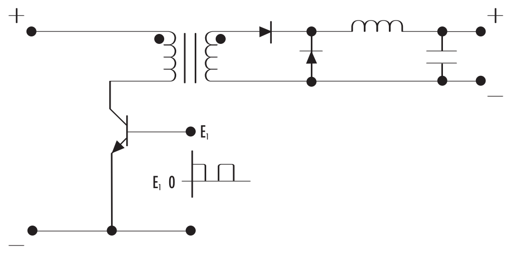
Above 20 kHz, core losses increase. To operate the SPS at higher frequencies, it is necessary to operate the core flux levels lower than ±2 kg. Figure 3 shows the reduction in flux levels for MAGNETICS "P" ferrite material necessary to maintain constant 100mW/cm^3 core losses at various frequencies,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

FIGURE 3



FEED FORWARD CIRCUIT

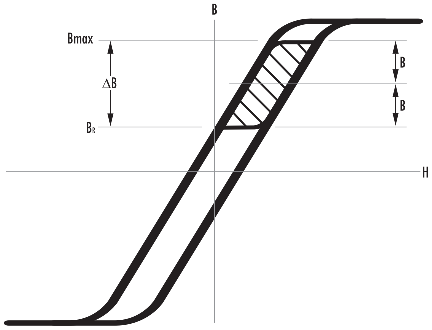
FIGURE 4A – TYPICAL FEED FORWARD SPS CIRCUIT



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.

Specific Circuit Examples

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

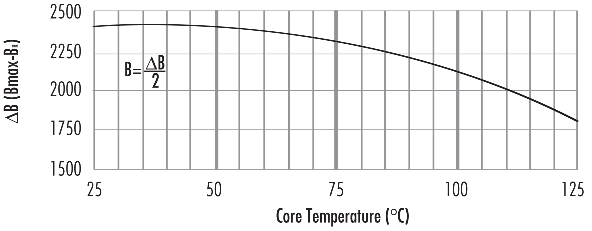


For ferrites used in this circuit, ΔB (or B max-B_R) is typically 2400 gauss or B (as applied to Equation 4) is ±1200 gauss as shown in Figure 4B. In the push-pull circuit, it was recommended that the peak flux density in the core should not exceed B = ±2000 gauss in order to keep core losses small. 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 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 form 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

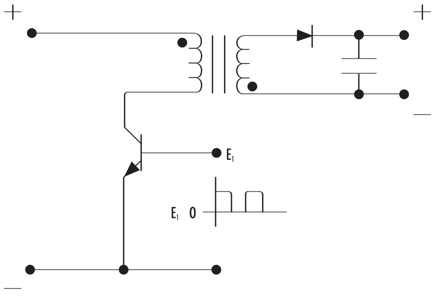
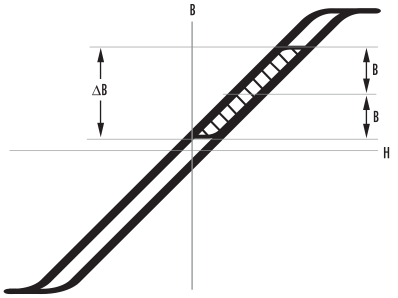


FIGURE 6B
HYSTERESIS LOOP OF MAGNETIC CORE IN FLYBACK CIRCUIT



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 be done using Equation 4. 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 required.

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. 13, pg 14.4.)

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

WaAc = Product of window area and core area (cm^4)
Po = Power Out (watts)
Dcma = 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.

Area Product Distribution (WaAc*)

TABLE 3 – FERRITE CORE SELECTION BY AREA PRODUCT DISTRIBUTION

WaAc* (cm^4)	**PC**	**RS,DS,HS**	**RM, EP**	**RM SOLID**	**PQ**	**EE LAM**	**EE,EEM,EFD**	**EE,EI PLANAR**	**UU, UI**	**ETD, EER**	**EC**	**TC**
See Section	6	7	8/9	8	10	11	11	11	11	12	12	13
0.001	40704							41309 (EE)				40601
0.002	40905		40707 (EP)				40904 40906					40603
0.004												
0.007	41107		41110(RM)									40705
0.010		41408 (RS,DS)	41010(EP)			41203			41106 (UI)		41003	41005
0.020	41408		41510(RM) 41313(EP)	41510		41205	41208 41209 41515 41707		41106(UU)		40907	41303
0.040			41812(RM)	41812			41709 42110					41206 41305
0.070	41811	42311 (RS,DS,HS)	41717(EP)		42610	41808						41306 41605
0.100	42213	42318 (HS)	42316(RM)	42316	42016 42614	41810 42510		42216(EE)				
0.200	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
0.400		43019 (RS,DS,HS)		42819	42625	42520	42515 43007	43618(EE) 43208(EE)				42207
0.700	43019		43723(RM)		43220	43515	43013		42220(UU) 42512(UU) 42515(UU)	43517		42507
1.00	43622	43622 (RS,DS,HS)		43723	43230	44317	43520 43524 44011	44308(EI)	42530(UU)	44119	43434 43521(EER)	42908
2.00	44229 44529	44229 (RS,DS,HS)			43535	44721	44020 44924	44308(EE) 45810(EI)	44119(UU) 44121(UU)	45224 44216(EER)	43939 43615 44444 45032	43610 43813
4.00					44040	45724	44022 45021	46410(EI)	44125(UU) 44130(UU)	44949	44416	
7.00							45528 46016	45810(EE) 46409(EE)				
10.00							45530	46410(EE)		47035 47228		44916 44925 46113
20.00							48020				47054	47313 47325
40.00								49938(EE)				48613
100							49928		49925(UU) 49925(UI)			

*Bobbin window and core area product. For bobbins other than those in this catalog, WaAc may need to be recalculated.

Typical Power Handling

TABLE 4 – FERRITE CORE SELECTION LISTED BY TYPICAL POWER HANDLING CAPABILITIES (WATTS) (F, P AND R MATERIALS) (FOR PUSH-PULL SQUARE WAVE OPERATIONS, SEE NOTES BELOW)

WATTAGE											
@F= 20KHZ	**@F= 50KHZ**	**@F= 100KHZ**	**@F= 250KHZ**	**POT-RS-RM CORES**	**DS CORES**	**EP CORES**	**PQ CORES**	**E-CORES**	**LOW-PROFILE PLANAR CORES**	**EC-ETD U CORES**	**TC TOROIDS**
See Section				6/7/8	7	9	10	11	11	12	13
2	3	4	7	41408-PC		41313		41707	41709 42107 42110		41206 41303
5	8	11	21	41811-PC 42311-RS 42809-RM	42311	41717		41808	42610-PQ 42216-EC		41306 41605
12	18	27	53	42316-RM			42016	41810, 42211	42614-PQ		
13	20	30	59					42510			
15	22	32	62	42213-PC							
18	28	43	84	42318-RS	42318		42020		43618-E, I		42106
19	30	48	94		42616	42120			43208-E, I 44008-E, I		41809
26	42	58	113					42810, 42520			42206
28	45	65	127	42819-RM				42515			42109
30	49	70	137	42616-PC			42620				42207
33	53	80	156		43019				43618-EC		
40	61	95	185	43019-RS				43007	44008-EC		43205
42	70	100	195				42625		43208-EC		
48	75	110	215					43013			42212, 42507
60	100	150	293	43019-PC 43723-RM			43220	42530, 43009 43515 (E375)		43517 (EC35)	
70	110	170	332		43622				44308-E, I	43434 (ETD34)	42908
105	160	235	460					44011 (E40)			
110	190	250	480	43622-PC			43230				
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						44308-EC	43939 (ETD39)	
190	300	470	917		44229						43610
200	310	500	975					44721 (E625)		45032	
220	350	530	1,034				43535				43813
230	350	550	1,073					44020 (42/15)		44216	
260	400	600	1,170								43615
280	430	650	1,268	44229-PC				45021 (E50) 44924		45224 (EC52)	
300	450	700	1,365	44529-PC				44022 (42/20)	45810-EC	44444 (ETD44)	
340	550	850	1,658				44040				
360	580	870	1,697								43825
410	650	1,000	1,950					45724 (E75)	46410-E, I	44949 (ETD49)	44416
550	800	1,300	2,535					45528 (55/21) 46016 (E60)	45810-EC		44715
650	1,000	1,600	3,120								44916 44920
700	1,100	1,800	3,510					45530 (55/25)	46409-EC		
850	1,300	1,900	3,705						46410-EC		44925
900	1,500	2,000	3,900							47035 (EC70)	
1,000	1,600	2,500	4,875							45959 (ETD59)	46113
1,000	1,700	2,700	5,265					47228			
1,400	2,500	3,200	6,240								44932
1,600	2,600	3,700	7,215								47313
2,000	3,000	4,600	8,970					48020		47054	
2,800	4,200	6,500	12,675						49938-EC		48613
11,700	19,000	26,500	51,500							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/cm^3,
B Levels Used in this Chart are:@ 20kHz-2000 gauss @ 50kHz-1300 gauss @ 100kHz-900 gauss @ 250kHz-700 gauss.

SEE PAGE 4.7 — Area Product Distribution

Considerations

TEMPERATURE CONSIDERATIONS

The power handling ability of a ferrite transformer is limited by either the saturation of the core material or, more commonly, the temperature rise. Core material saturation is the limiting factor when the operating frequency is below 20kHz. Above this frequency temperature rise becomes the limitation.

Temperature rise is important for overall circuit reliability. Staying below a given temperature insures that wire insulation is valid, that nearby active components do not go beyond their rated temperature, and overall temperature requirements are met. Temperature rise is also very important for the core material point of view. As core temperature rises, core losses can rise and the maximum saturation flux density decreases. Thermal runaway can occur causing the core to heat up to its Curie temperature resulting in a loss of all magnetic properties and catastrophic failure. Newer ferrite power materials, like P and R material, attempt to mitigate this problem by being tailored to have decreasing losses to temperature of 70°C and 100°C respectively.

CORE LOSS—One of the two major factors effecting temperature rise is core loss. In a transformer, core loss is a function of the voltage applied across the primary winding. In an inductor core, it is a function of the varying current applied through the inductor. In either case the operating flux density level, or B level, needs to be determined to estimate the core loss. With the frequency and B level known, core loss can be estimated from the material core loss curves. A material loss density of 100mw/cm^3 is a common operating point generating about a 40°C temperature rise. Operating at levels of 200 or 300 mw/cm^3 can also be achieved, although forced air or heat sinks may need to be used.

WINDING CONSIDERATIONS—Copper loss is the second major contributor to temperature rise. Wire tables can be used as a guide to estimate an approximate wire size but final wire size is dependent on how hot the designer allows the wire to get. Magnet wire is commonly used and high frequency copper loss needs to be considered. Skin effects causes current to flow primarily on the surface of the wire. To combat this, multiple strands of magnet wire, which have a greater surface area compared to a single heavier gauge, are used. Stranded wire is also easier to wind particularly on toroids. Other wire alternatives, which increase surface areas, are foil and litz wire. Foil winding allows a very high current density. Foil should not be used in a core structure with significant air gap since excessive eddy currents would be present in the foil. Litz wire is very fine wire bundled together. It is similar to stranded wire except the wire is woven to allow each strand to alternate between the outside and the inside of the bundle over a given length.

CORE GEOMETRY—The core shape also affects temperature and those that dissipate heat well are desirable. E core shapes dissipate heat well. Toroids, along with power shapes like the PQ, are satisfactory. Older telecommunication shapes, such as pot cores or RM cores, do a poor job of dissipating heat but do offer shielding advantages. Newer shapes, such as planar cores, offer a large flat surface ideal for attachment of a heat sink.

TRANSFORMER EQUATIONS

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

$$Np = \frac{V_p \times 10^8}{4BAf} \qquad Ns = \frac{V_s}{V_p} Np$$

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

$$KWa = N_pA_{wp} = N_sA_{ws}$$

Where

A_{wp} = primary wire area $\qquad A_{ws}$ = secondary wire area
Assume K = 0.40 for toroids; 0.60 for pot cores and E-U-I cores
Assume $N_pA_{wp} = 1.1\ N_sA_{ws}$ to allow for losses and feedback winding

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

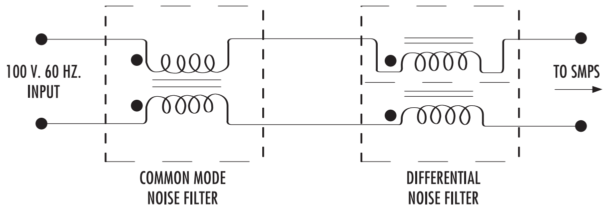
$$\text{Voltage Regulations (\%)} = \frac{R_s + (N_s/N_p)^2R_p}{R_{load}} \times 100$$

INDUCTOR CORE SELECTION

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 can be in series, or in many cases, the common mode filter is used alone.

FIGURE 8



INDUCTOR CORE SELECTION CONT...

COMMON MODE FILTER

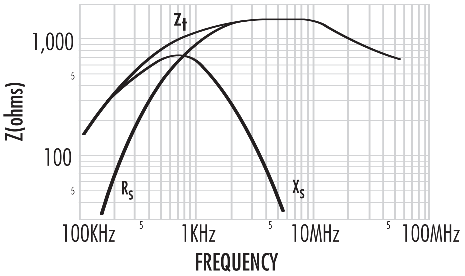
In a CMN filter, each winding of the inductor is connected in series with one of the input power lines. The connections and phasing of the inductor windings are such that flux created by one winding cancels the flux of the second winding. The insertion impedance of the inductor to the input power line 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, J42206TC 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 of 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, cause the resistance to dominate the source of impedance at high frequencies.

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

FIGURE 10

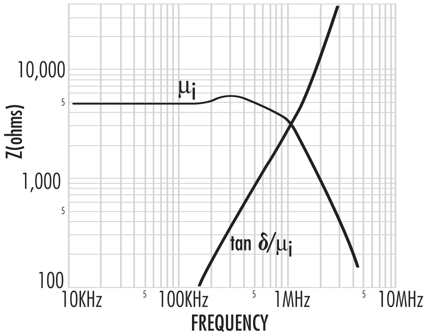
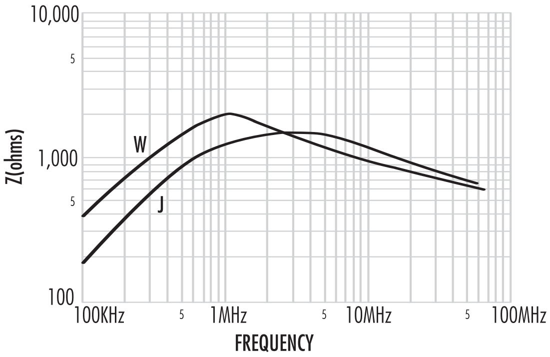


Figure 11 shows total impedance vs. frequency for two different materials. J material has a high total impedance over the range of 1 to 20MHz. It is most widely used for common mode filter chokes. Under 1MHz, W material has 20-50% more impedance than J. It is often used in place of J when low frequency noise if the major problem. 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.

Inductor Design

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/cm^2 does not significantly heat up the wire. A more aggressive 800 amps/cm^2 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/cm^2:

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

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

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

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/cm^2, 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 W41809TC. Page 13.6 lists the core sizes and A_L values. Using an A_L of 12,200 mH/1,000 turns, equation 3 yields N = 12 turns per side. Using 800 amps/cm^2, equation 2 yields AWG = 21.

FIGURE 12: COMMON MODE INDUCTOR WINDING ARRANGEMENT

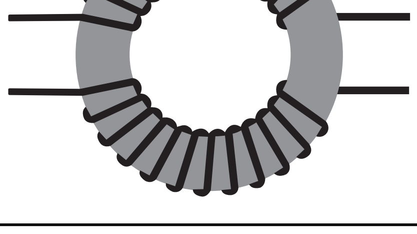
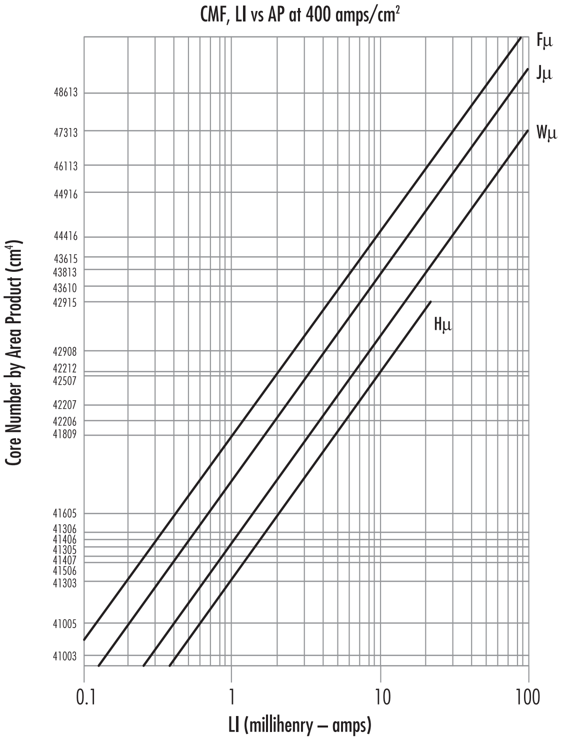
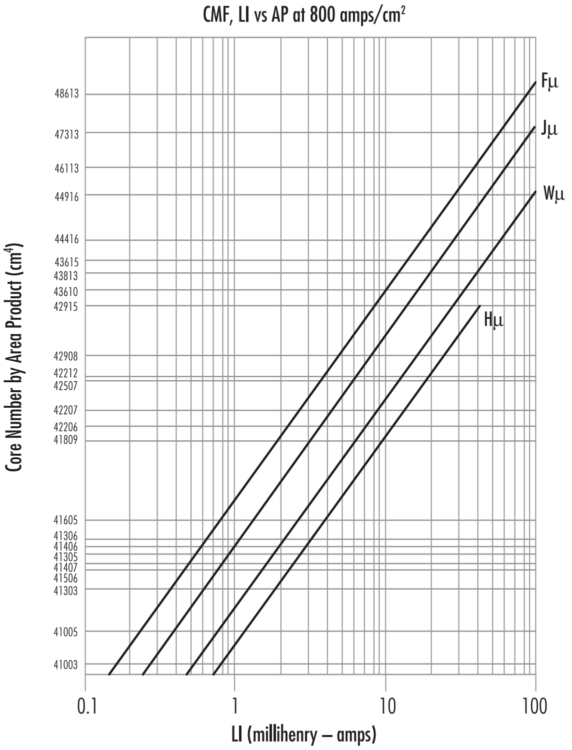


FIG. 13: CORE SELECTION AT 400 amps/cm^2



CMF, LI vs AP at 400 amps/cm²

FIG. 14: CORE SELECTION AT 800 amps/cm^2



CMF, LI vs AP at 800 amps/cm²

Inductor Design

HALL EFFECT DEVICES

Edwin H. Hall observed the "Hall Effect" phenomenon at John 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 flow in a wire, the wire is wrapped around a slotted ferromagnetic core, creating an electromagnet. 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:

$$\mu_e = \frac{b\ell_e}{.4\pi NI} \qquad (1)$$

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

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

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

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.

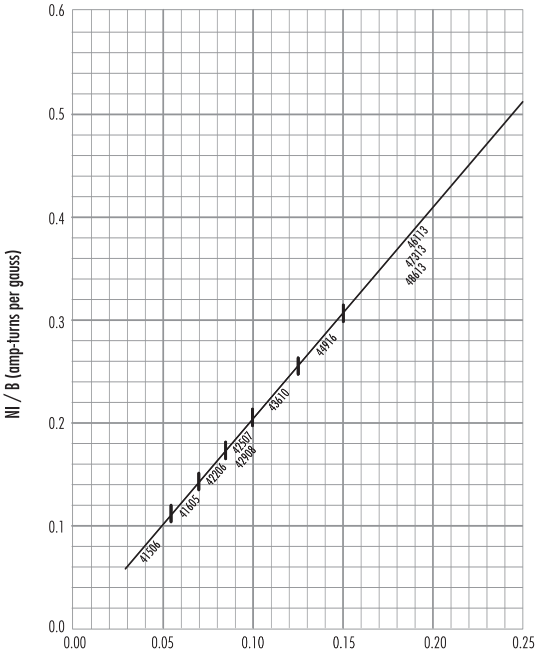
2. Using Figure 15, follow the NI/B value from the vertical axis to the diagonal line to choose a ferrite core size. Drop down from the diagonal line to the horizontal axis to determine the gap length. The core sizes indicated on the selector chart take into account gap length versus cross-section dimensions 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.

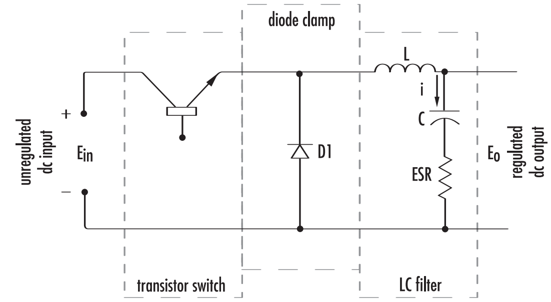
Diamond bonded-tool machining is the preferred method of cutting ferrite. The bonded diamond particle size should be approximately 100 to 170 mesh (150 to 90 μm). The peripheral speed of the cutting wheel should be 5,000 to 6,000 feet/minute (1,500 to 1,800 meters/minute). The depth of the cut may be 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.

GAPPED TOROID SELECTOR CHART



Inductor Design

FIG. 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 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

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

Finally, the maximum ESR of the capacitor is

$$ESR\ max = \Delta e_o/\Delta_i \qquad (7)$$

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 two design applications 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 pgs 4.15–4.18. 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 of 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 pg 5.8 using 500 circular mils per amp.

Inductor Design

EXAMPLE

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

E_o $=5$ volts
Δe_o $=0.50$ 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.

   $L = 5(4.3 \times 10^{-5})/2 = 0.107$ millihenries

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

   $C = 2/8\ (18{,}700)\ (0.50) = 26.7$ μ farads
   and ESR max $= 0.50/2 = .25$ ohms

5. The product of $LI^2 = (0.107)\ (8)^2 = 6.9$ millijoules

6. Due to the many shapes available in ferrites, there can be several choices for the selection. 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_L315
   (b) 44229 solid center post core, A_L315
   (c) 43622 pot core, A_L400
   (d) 43230 PQ core, A_L250

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

A_L	Turns
250	21
315	19
400	17

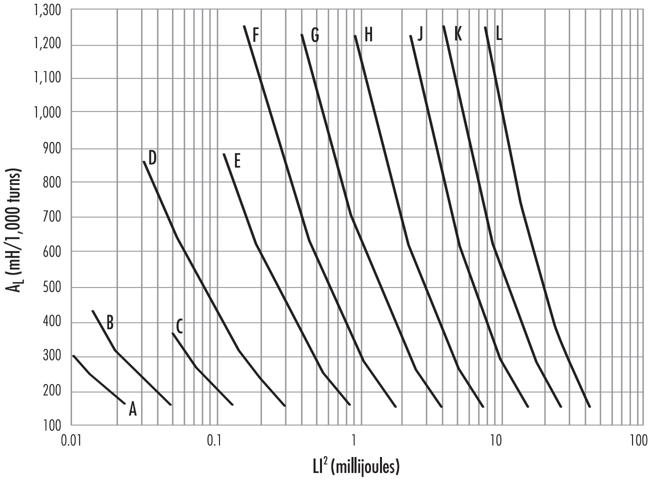
8. Use #14 wire

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 Magnetics Powder Core Catalog and Brochure SR-IA, "Inductor Design in Switching Regulators."

FOR REFERENCES, SEE PAGE 14.4

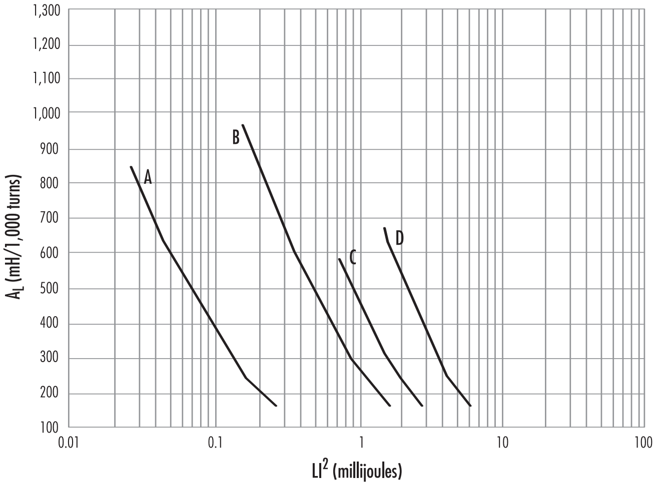
Selector Charts

PC (POT) CORES



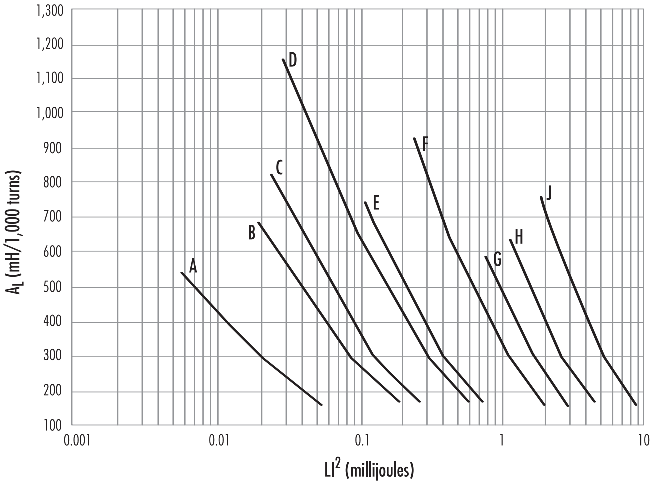
A — 40704
B — 40905
C — 41107
D — 41408
E — 41811
F — 42213
G — 42616
H — 43019
J — 43622
K — 44229
L — 44529

RS (ROUND-SLAB) & DS (DOUBLE-SLAB) CORES



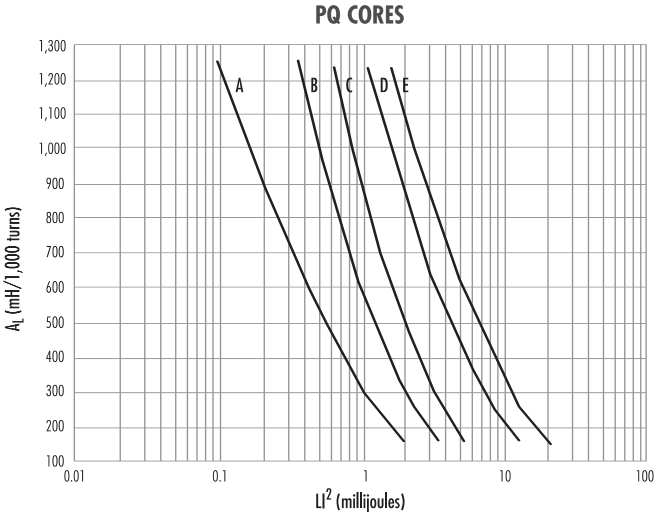
A — 41408 (RS)
B — 42311 (DS, RS)
42318 (DS, RS)
C — 42616 (DS)
D — 43019 (DS, RS)

RM AND EP CORES

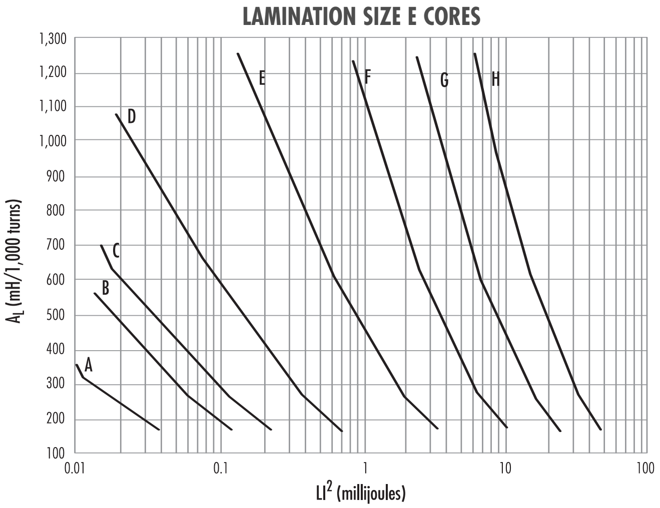


A — 40707 (EP7)
41010 (EP10)
41110 (RM4)
B — 41313 (EP13)
C — 41510 (RM5)
D — 41717 (EP17)
E — 41812 (RM6)
F — 42316 (RM8)
G — 42120 (EP20)
H — 42819 (RM10)
J — N43723 (RM12)

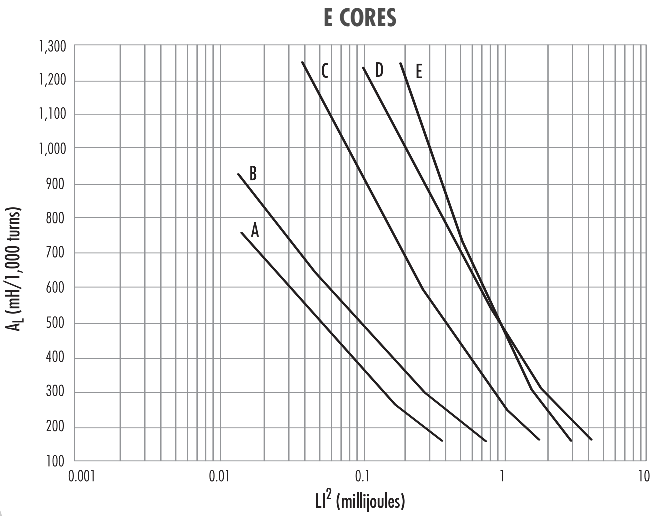
Selector Charts



A — 42016
42020
B — 42614
C — 42610
42620
42625
43214
D — 43220
43230
E — 43535
44040

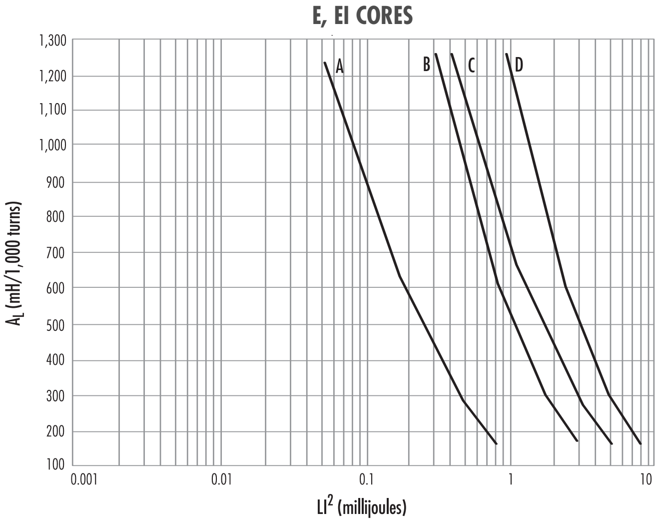


A — 41203 (EE)
B — 41707 (EE)
C — 41808 (EE)
D — 42510 (EE)
E — 43009 (EE)
43515 (EE)
F — 44317 (EE)
G — 44721 (EE)
H — 45724 (EE)

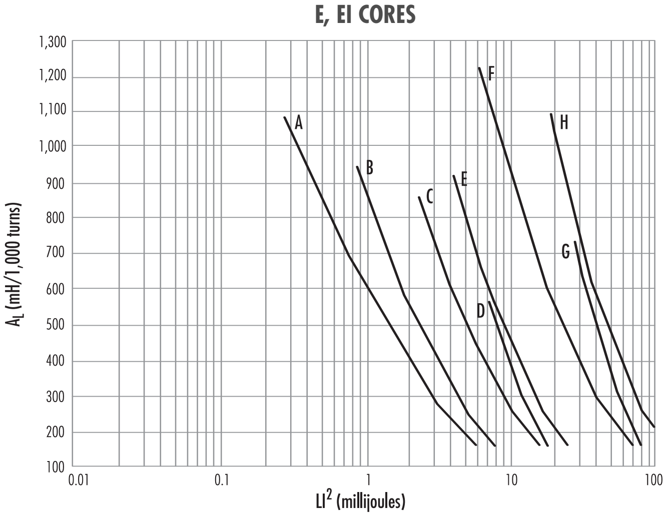


A — 41205 (EE)
B — 42515 (EE)
C — 41810 (EE)
43007 (EE)
D — 42530 (EE)
43520 (EE)
E — 42520 (EE)

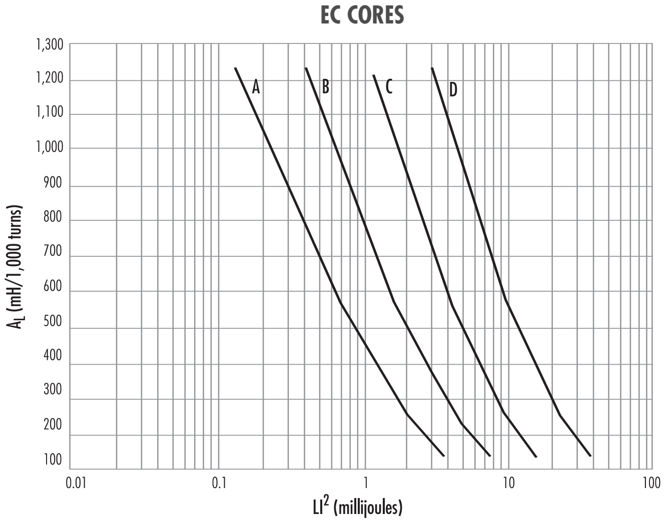
Selector Charts



A — 41805 (EE, EI)
B — 42216 (EE, EI)
C — 44008 (EE, EI)
D — 43618 (EE, EI)

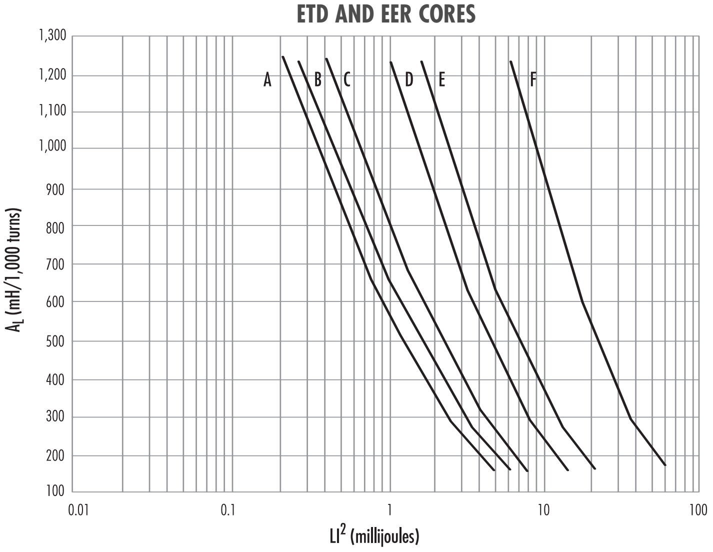


A — 44016 (EE)
B — 44011 (EE)
C — 44020 (EE)
D — 44308 (EE, EI)
E — 44022 (EE)
46016 (EE)
F — 45528 (EE)
45530 (EE)
47228 (EE)
48020 (EE)
G — 46410 (EE)
H — 49938 (EE, EI)

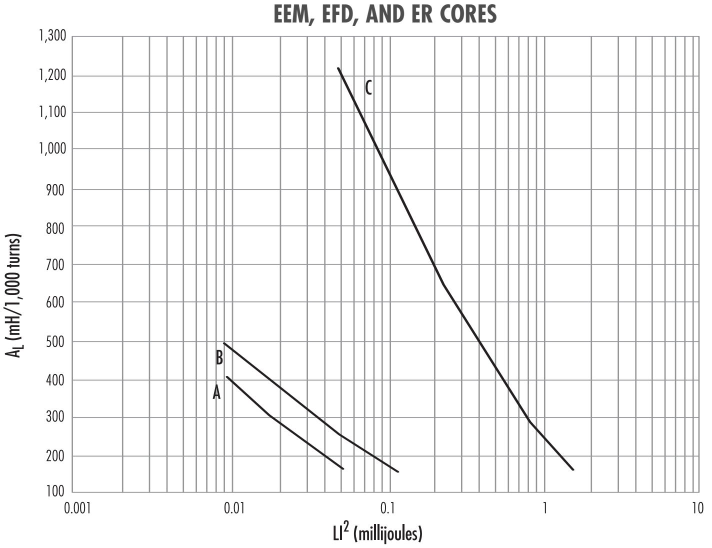


A — 43517
B — 44119
C — 45224
D — 47035

Selector Charts



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

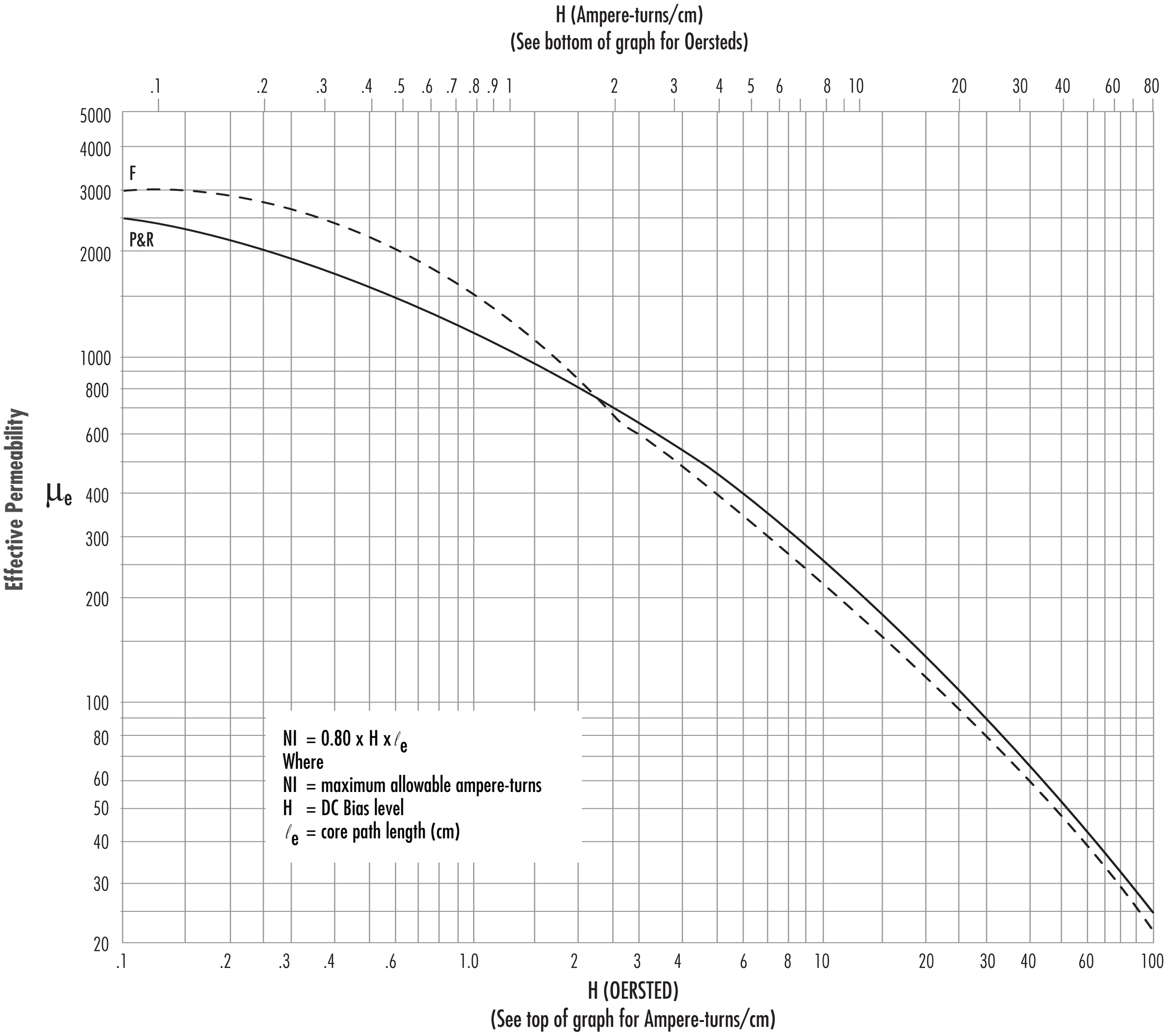


A — 40906 (ER 9.5)
B — 41515 (EFD15)
C — 42523 (EFD25)

DC Bias Data

DC BIAS DATA — FOR GAPPED APPLICATIONS

μ_e vs. H



NI = 0.80 x H x ℓ_e
Where
NI = maximum allowable ampere-turns
H = DC Bias level
ℓ_e = core path length (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 R42213A315 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) = 0.80 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 (cm^2)
A_L = inductance/1,000 turns (mH)
μ_i = initial permeability
ℓ_g = gap length (cm)