POWER FACTOR CORRECTION INDUCTOR DESIGN FOR SWITCHED MODE POWER SUPPLIES USING

metglas® POWERLITE®

C-Cores

INTRODUCTION

Since the mid 1970's, the power processing industry has been continuously discussing the need to correct input power factor for ac line driven equipment. Power Factor by definition in a single-phase ac system is the ratio of the input power in watts (V I cos Φ, where Φ is the phase angle between V and 1) to the input in volt-amps (VI) i.e.

POWER FACTOR = (VI cos Φ) / VI=cos Φ

For a purely resistive load, the input voltage and current are always in phase; therefore, the power factor is unity. This is why a power factor correction circuit is sometimes referred to as a resistor emulator. In a purely reactive load, the input voltage and current are always 90° out of phase; therefore, the power factor is zero.

When we examine the input current waveform to off-line equipment that utilizes a rectifier/capacitor input filter, we observe that the input current is supplied in narrow pulses at the peak of the input voltage waveform. This occurs because the input capacitor only draws current from the line when its voltage has fallen below that of the rectified line voltage. This results in off-line equipment with a low power factor, and a correspondingly high harmonic current content. These currents cause unwanted line noise and do not effectively supply power to the equipment.

The fundamental rms line current in non-power factor corrected equipment utilizing a rectifier/capacitor input filter can be as much as two times that for equipment which is power factor corrected. As a result, full utilization of the line cannot be achieved with non-power factor corrected equipment without increasing the rating of the line circuit breaker above that of the line current.

In light of this new requirement, equipment manufacturers are faced with several problems:

** The task of developing a solution that will limit the harmonic currents below that which is required by the IEC 555 standard.

** The associated costs of implementing this solution.

** The additional size and weight associated with this solution.

The first difficulty has been addressed for a number of years, with several topologies that provide the required harmonic current limitation. The most notable and widespread of these is the continuous mode boost pre-regulator. However, irrespective of which topology is used, the remaining problems still exist: the additional cost, size and weight of the equipment. In a discipline where cost, size and weight reduction have been the significant driving forces over the past two decades, this surely represents a disappointing step backwards.

Common to all the topologies being used for harmonic current limitation is an inductive component. When made of ferrite, MPP or powder iron materials, it represents a significant portion of the additional cost, size and weight of the overall solution.

With this and many more applications in mind, Honeywell has developed a new product line, POWERLITE C-Cores using amorphous alloy technology to address the problems of cost, size and weight. Utilizing the high saturation flux density and low frequency dependent losses of amorphous Alloy-SA1 the inductive component can be made smaller and lighter. At present (for a given performance) the amorphous solution is likely to be lower in cost at power levels above 800W - 1000W It is expected that this power level will extend down to the lower threshold of the standard in the near future.

The following pages describe a step-by-step procedure for designing the inductive element in a continuous mode boost pre-regulator for correcting input power factor and limiting harmonic currents.

OPERATION OF THE BOOST PRE-REGULATOR

The ideal boost switch mode power converter is shown in Figure 1. During the on-time of the switch, the inductor current ramps upwards from its initial value I_a towards its final value I_b just before the switch turns off. After the switch turns off, the energy stored in the inductor during the on time of the switch is delivered to the output network, and the inductor current ramps downwards from I_b If the converter is operating in the continuous mode, the current ramps down to a value which is greater than zero.

As mentioned earlier, the output voltage of the boost converter must be higher than the input voltage. If we choose an output voltage that is higher than the peak ac line voltage, the converter will operate over the full input voltage range (from zero volts to the ac line peak, approx. 380 V). This enables universal input to the converter, without the need for additional range-switching circuitry

In order for the inductor current to be continuous, the ripple current Al in the inductor should not exceed two times the minimum input peak current. The inductance value which guarantees this condition is the critical inductance L_{crit}' and is given by the expression
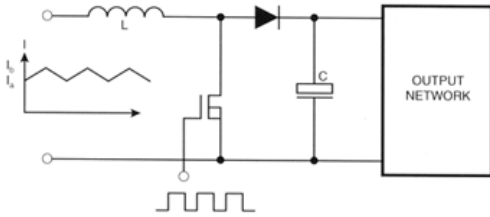
$$L_{crit} = \frac{V_{in\,min}\left(1 - V_{in\,min}/V_{out}\right)}{\Delta I\, f_{sw}}$$

L_{crit} …….. Critical Inductance (H)
$V_{in\,min}$ ….. Minimum Input dc Voltage (V)
V_{out} …….. Output dc Voltage (V)
ΔI ……… Ripple Current (A)
f_{sw} ………… Switching Frequency (kHz)

Where $V_{in\,(dc)min} = \sqrt{2}\,(V_{in\,(ac)\,min})$

Having Calculated the inductance to maintain the required ΔI, the peak current that the inductor must conduct without saturating can be calculated.

$$I_{in(rms)\,max} = P_{in\,max} / V_{in(rms)\,min}$$
$$= P_{out\,max} / (V_{in(rms)\,min}\,\eta_{PREG})$$

There fore there is a sinusoidal input current

$$I_{in\,pk} = \sqrt{2}\,(I_{in(rms)\,max}) + \Delta I / 2$$

Overload current conditions should be taken into account.

Now that the necessary inductor parameters L_{crit} and $I_{in\,pk}$ have been calculated, we can continue with the design procedure for the power factor correction inductor.

FIGURE 1

Many of the properties of the boost converter can be determined by examining the schematic. The output volts must be higher than the input voltage; if this were not true the inductor would not pass current to the output network of the converter. In the continuous mode, the value of the ripple current in the inductor can be made arbitrarily small by increasing the value of the inductance.

If we modify the dc-dc boost converter by including a bridge rectifier at the input, we obtain the circuit shown in. Figure 2. This is the typical power train for a high power factor boost pre-regulator. By suitably controlling the duty cycle of the switch, this converter will draw a sinusoidal current from an ac source that is in phase with the ac source voltage, thereby creating a converter with a unity power factor.
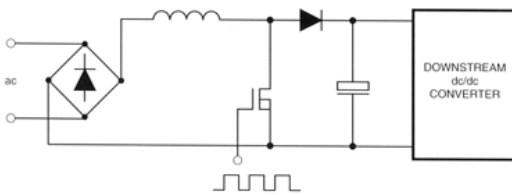

FIGURE 2

DESIGN PROCEDURE	DESIGN EXAMPLE

Step #1 Determine the input data

Topology		*Continuous Mode Boost Pre-regulator*	
Output Power (overload)	P_{out}	Output Power (overload)	**2200W**
Switching Frequency	f_{sw}	Switching Frequency	**50,000 Hz**
Output dc Voltage	V_{out}	Output dc Voltage	**380 V**
Minimum Input as Voltage	$V_{in (rms) main}$	Minimum Input as Voltage	**90 V**
Maximum Input as Voltage	$V_{in (rms) max}$	Maximum Input as Voltage	**260 V**
Maximum Temp Above Ambient	°C	Maximum Temp Above Ambient	**50 °C**
Efficiency of the Inductor	NL	Efficiency of the Inductor	**99%**
Efficiency of the pre-regulator	N_{preg}	Efficiency of the pre-regulator	**95%**

Step #2 Estimate Ripple Current and Critical Inductance

Assuming the inductor efficiency is 99%

$$\mathbf{P_{tot} = (1 - \eta_L) (P_{out} / \eta_{preg})} \quad \text{Watts}$$

$$P_{tot} = (1 - 0.99) (2200 / 0.95) = \mathbf{23.16\ W}$$

For a theoretically optimized design, the core and copper losses are:

$$\mathbf{P_{core} = P_{cu} = P_{tot} / 2} \quad \mathbf{Watts}$$

$$P_{core} = P_{cu} = 23.16 / 2 = \mathbf{11.58\ W}$$

Select a core from **Table 1** based on total losses anticipated and the required maximum $\Delta T° C$

22 W dissipated in an **AMCC-25** will give rise to $\Delta T \approx \mathbf{50°C}$

Calculate power loss for the core weight:

$$\mathbf{P = P_{tot} / (2\ wt)} \quad \text{Watts/kg}$$

$$P = 23.16 / [2\ (0.38)] = \mathbf{30.47\ W/kg}$$

Determine $\mathbf{B_{ac}}$ that will give rise to P at the switching frequency f_{sw}. For Honeywell **POWERLITE**™ C-Cores , we have:

$$\mathbf{P = 6.5\ f_{sw}{}^{1.51}\ B_{ac}{}^{1.74}} \quad \text{Watts/kg}$$

$$\mathbf{B_{ac} = [P / (6.5\ f_{sw}{}^{1.51})]^{1/1.74}} \quad \text{Tesla}$$

$$B_{ac} = [30.47 / ((6.5)\ 50^{1.51})]^{1/1.74} = \mathbf{0.0815\ T}$$

$$\mathbf{\Delta B = 2\ B_{ac}} \quad \text{Tesla}$$

Assuming the peak line frequency current corresponds to approximately 1.4 Tesla the ripple current is:

$$\Delta I = \Delta B\ /\ B\ [\sqrt{2}\ (P_{out})\ /\ (\eta_{preg}\ V_{in(ac)\ min})]\quad \text{Amps}$$

$$\Delta I = (0.163\ /\ 1.4)\ [\sqrt{2}\ (2200)\ /\ ((0.95)\ (90))]$$
$$= 4.23\ A_{pk\text{-}pk}$$

Calculate the critical inductance based on ΔI $L_{crit} =$

$$\frac{\sqrt{2}\ V_{in(ac)\ min}\ [1-((\sqrt{2}\ V_{in(ac)\ min})/V_{out})]}{\Delta I\ f_{sw}}$$
$$\text{Henry}$$

$$L_{crit} = \frac{\sqrt{2}\ (90)\ [1-(\sqrt{2}\ (90\ /\ 380))}{(4.23)\ (50,000)} = 400\mu H$$

Approximate Power Dissipation in Watts for a Given Temperature Rise °C

CORE	SA cm²	20°C	25°C	30°C	35°C	40°C	45°C	50°C	CORE	SA cm²	20°C	25°C	30°C	35°C	40°C	45°C	50°C
AMCC-6.3	103.4	3.8	4.9	6.1	7.4	8.7	10.0	11.3	AMCC-100	373.0	13.6	17.8	22.1	26.6	31.3	36.0	41.0
AMCC-8	115.9	4.2	5.5	6.9	8.3	9.7	11.2	12.7	AMCC-125	464.0	16.9	22.1	27.5	33.1	38.9	44.8	51.0
AMCC-10	132.1	4.8	6.3	7.8	9.4	11.0	12.7	14.5	AMCC-160	485.0	17.7	23.1	28.8	34.6	40.6	46.8	53.0
AMCC-16A	143.1	5.2	6.8	8.5	10.2	12.0	13.8	15.7	AMCC-200	526.0	19.2	25.1	31.2	37.5	44.0	51.0	57.0
AMCC-16B	160.3	5.8	7.6	9.5	11.4	13.4	15.5	17.6	AMCC-250	592.0	21.6	28.2	35.1	42.3	49.6	57.1	65.0
AMCC-20	172.3	6.3	8.2	10.2	12.3	14.4	16.6	18.9	AMCC-320	707.0	25.8	33.7	41.9	50.5	59.2	68.2	77.5
AMCC-25	202.2	7.4	9.6	12	14.4	16.9	19.5	22.2	AMCC-400	780.0	28.4	37.2	46.3	55.7	65.4	75.3	85.5
AMCC-32	216.0	7.9	10.3	12.8	15.4	18.1	20.8	23.7	AMCC-500	854.5	31.2	40.7	50.7	61.0	71.6	82.5	93.6
AMCC-40	230.0	8.4	11.0	13.6	16.4	19.3	22.2	25.2	AMCC-630	934.0	34.1	44.5	55.4	66.7	78.3	90.2	102.0
AMCC-50	303.5	11.1	14.5	18.0	21.7	25.4	29.3	33.2	AMCC-800A	1013	36.9	48.3	60.1	72.3	84.9	97.8	111.0
AMCC-63	321.0	11.7	15.3	19.0	22.9	26.9	31.0	35.1	AMCC-800B	1171	42.7	55.8	69.5	83.6	98.1	113.0	128.0
AMCC-80	356.0	13.0	17.0	21.1	25.4	29.8	34.3	39.0	AMCC-1000	1292	47.1	61.6	76.6	92.2	108.3	124.7	142.0

Table 1

Step #3 Calculate the Energy Storage Requirement , E

Having calculated L_{crit} and I_{in} (see Operation of Boost Pre-Regulator- p2) , we can now calculate the energy storage requirements of the inductor.

$$E = 0.5\ (L_{crit}\ I_{in\ pk}^{2})\quad \text{Joules}$$

$$E = 0.5\ (400 \times 10^{-6})\ (38.5)^{2} = 0.296\ \text{Joules}$$

$$I_{in\ pk} = \sqrt{2}\ P_{out}/(\eta_{PREG}\ V_{in(ac)\ min}) + \Delta I/2\quad \text{Amps}$$

Step #4 Calculate the Required Core Area Product, W_aA_c *

$$W_aA_c = \frac{2\,E\,(10^4)}{B_{max}\,J\,K} \quad cm^4$$

$$W_aA_c = \frac{(2)\,(0.296)\,(10^4)}{(1.4)\,(500)\,(0.4)} = 21.17\ cm^4$$

E…………………Energy (Joules)
B_{max}……………Max Peak Flux Density (Tesla)
J…………………Current Density (A/cm^2)
K…………………Window Utilization Factor

This should agree to within a size of the core selected in Step #2

A complete derivation can be found in Appendix A.

Step #5 Select a Core

From the tabel of cores in Appendix B, Choose a core with W_aA_c greater than that calculated in Step #4.

a = 1.3 cm
b = 1.5 cm
c = 5.6 cm
d = 2.5 cm
e = 4.1 cm
f = 8.2 cm

ℓm = 19.6 cm
Ac = 2.7 cm^2
wt = 0.38 kg
W_aA_c = 22.7 cm^4
SA = 202.2 cm^2

Step #6 Calculate the Number of Turns Required, N

$$N = \frac{L_{crit}\,I_{in\ pk}\,(10^4)}{B_{max}\,A_c} \quad Turns$$

$$N = \frac{(400 \times 10^{-6})\,(38.5)\,(10^4)}{(1.4)\,(2.7)} = 41\ Turns$$

Step #7 Calculate the Air Gap Required, ℓg

$$L = \frac{0.4\,\pi\,N^2\,A_c\,(10^{-8})}{\ell_g + (\ell_m / \mu\Delta)} \quad Henry$$

ℓg………………… Total Air Gap (cm)
ℓm………………… Magnetic Path Length of core (cm)
$\mu\,\Delta$………………… Incremental Permeability at operating point on B-H loop = 1000

$$\ell_g = \frac{0.4\ \pi\ N^2\ A_c\ (10^{-8})}{L} - \frac{\ell_m}{\mu\Delta} \quad \text{cm}$$

From Step #6 we have:

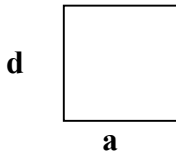$$\frac{N\ A_c}{L} = \frac{I_{in\ pk}\ (10^4)}{B_{max}}$$

Therefore:

$$\boxed{\ell_g = \frac{0.4\ \pi\ N\ I_{in\ pk}\ (10^{-4})}{B_{max}} - \frac{\ell_m}{\mu\Delta} \quad \text{cm}}$$

$$\boxed{\ell_g = \frac{0.4\ \pi\ (41)\ (38.5)\ (10^{-4})}{1.4} - \frac{19.6}{1000}}$$
$$= 0.12\ \text{cm} = 0.06\ \text{cm/leg}$$

Step #8 Estimate the Amount of Fringing Flux , F

For Honeywell **POWERLITE**™ C-Cores , the physical core cross section dimensions are given by **a** and **d**

d □

a

The ratio of the air cross section to the physical cross section of the core is approximated by:

$$\boxed{F = \frac{(a + \ell_g)\ (d + \ell_g)}{a\ d}}$$

$$\boxed{F = \frac{(1.3 + 0.06)\ (2.5 + 0.06)}{(1.3)\ (2.5)} = 1.07}$$

F is sometimes known as the Fringing Factor. The Fring Flux has the effect of reducing the reluctance of the gap by a factor of F, which is turn increases the inductance of the component ratio by a factor of F.

Step #9 Calculate the Corrected Turns to give the required Inductance Value

From the first equation in Step #7 we have:

$$\boxed{N = \frac{L_{crit}\ I_{in\ pk}\ (10^4)}{B_{max}\ A_c} \quad \text{Turns}}$$

Because of the fringing flux A_c is increased by the fringing factor F. Therefore:

$$N = \sqrt{\frac{L\left[\ell_g + (\ell_m/\mu\Delta)\right](10^8)}{0.4\,\pi\,A_c\,F}} \quad \text{Turns}$$

$$N = \sqrt{\frac{(400 \times 10^{-6})\,[0.12 + (19.6/1000)]\,(10^8)}{(0.4)\,\pi\,(2.7)\,(1.07)}}$$

$$= 39 \text{ Turns}$$

Step #10 Calculate the Conductor Area, A_x

If the conductor has a circular cross section then is is a reasonable assumption that a window utilization factor of 0.4 can be used. If the conductor has a rectulangular cross section then it is a reasonable assumption that a window utilization factor of up to 0.6 can be used. Both the above window utilization factors will leave sufficient sapce for a bobbin and insulation between windings.

For Honeywell POWERLITE C-Cores the window area is given byu the dimensions **b** and **c**.

c

b
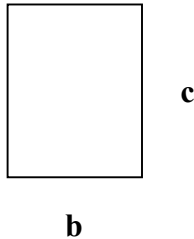
The coss sectional area of the conductor is :

$$A_x = \frac{b\,c\,K}{N} \quad \text{cm}^2$$

$$A_x = \frac{(1.5)\,(5.6)\,(0.4)}{39} = 0.086 \text{ cm}^2$$

At this point we can also calculate the resistance per unit length for the calculated cross section. For copper at a given temperature T, the resistivity is given by:

$$\rho = 1.724\,[1 + 0.0042\,(T - 20)]\,(10^{-6}) \quad \Omega\text{-cm}$$

If we assume that the maximum ambient temperature is 50°C , and the losses in the component give rise to an additional temperature rise of 30°C , then we should use a value of 80°C , when calculating the resistivity of copper. This responds to :

$$\rho = 2.16 \ \mu\Omega\text{-cm}$$

The resistance per unit length of conductor is given by:

$$R_{unit} = \rho/A_x$$

$$R_{unit} = \frac{2.16\,N}{b\,c\,K} \quad \mu\Omega\text{-cm}$$

$$R_{unit} = \frac{(2.16)\,(39)\,(10^{-6})}{(1.5)\,(5.6)\,(0.4)} = 25.0 \ \mu\Omega\text{-cm}$$

Step # 11 Calculate the copper Losses, P_{cu}

For Honeywell **POWERLITE**™ C-Cores , the mean turn
length is approximated by:

$$MTL = (2) (a + 2 b + d) \quad cm$$

$$MTL = (2) [1.3 + (2) (1.5) + 2.5] = 13.6 \text{ cm}$$

The total resistance for the winding is given by:

$$\Omega_{tot} = R_{unit} \text{ MTL N} \quad \Omega$$

$$\Omega_{tot} = (25.0 \times 10^{-6}) (13.6) (39) = 13.26 \text{ m}\Omega$$

If the current has a significant high frequency component
then the resistance of the winding will increase due to the
skin effect. The skin effectin high frequency designs can be
minimized by using Litx or foil windings.

The copper lossed can be calculated:

$$P_{cu} = I^2 \text{ rms } \Omega \text{ tot}   \text{ Watts}$$

$$P_{cu} = (26)^2 (13.26)(10^{-3)} = \textbf{8.96 W}$$

Appendix C demonstrates how to calculate the rms current
of a complex wave form.

Step #12 Calculate the ac Flux in the GAP (and Core)

$$B_{ac} = \frac{0.4 \, \pi \, N \, (\Delta I / 2) \, (10^{-4})}{l_g} \quad Tesla$$

$$B_{ac} = \frac{0.4 \, \pi \, 39 \, (4.23 / 2) \, (10^{-4})}{0.12} = \textbf{0.086 T}$$

Step #13 Calculate the Core Losses

For Honeywell POWERLITE C-Cores, recall that the losses
are given by the expressions:

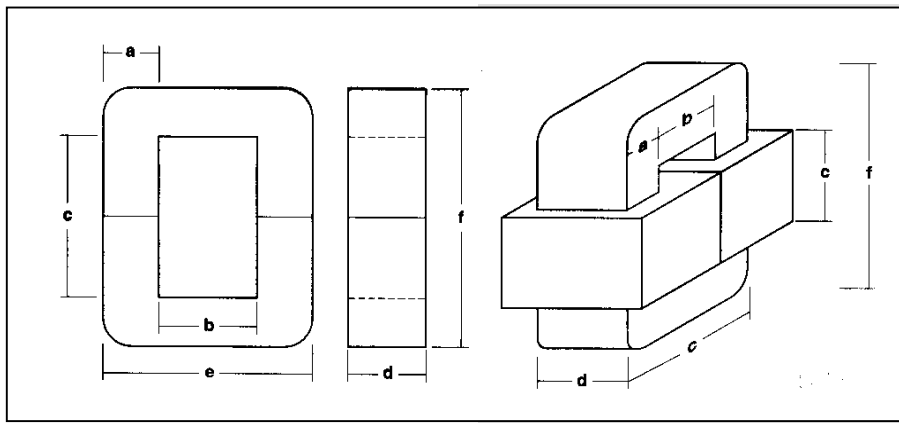$$P_{core} = 6.5 \, f_{sw}^{1.51} \, B_{ac}^{1.74} \quad Watts/kg$$

$$P_{core} = 6.5 \, f_{sw}^{1.51} \, B_{ac}^{1.74} \, wt \quad Watts$$

$$P_{core} = 6.5 \, (50)^{1.51} \, (0.086)^{1.74} \, (0.38) = \textbf{12.71 W}$$

Note: f_{sw} in kHz

Step #14 Estimate the Convective Surface Area

Painstaking accuracy can be employed to calculate the actual surface are of the wound component. However the boundry layer associated with the natural convection in the air tends to round off the surface contours and so the effective convective surface area is approximated by the surface area of a box barely enclosing the wound component.

For Honeywell POWERLITE C-Cores, this area is calculated as follows:

$$SA = 2f(b+d) + 2(b+d)(b+e) + 2f(b+e) \quad cm^2$$

$$SA = 2(8.2)(1.5 + 2.5)$$
$$+ 2(1.5 + 2.5)(1.5 + 4.1)$$
$$+ 2(8.2)(1.5 + 4.1) = 202.2 \ cm^2$$

Step #15 Calculate the Approximate Temperature Rise

The temperature rise of the wound component can be apporximated by using the formula below:

$$\mathbf{P_{tot} = P_{core} + P_{cu}} \quad \textbf{Watts}$$

$$\mathbf{\Delta T = (P_{tot}/SA)^{0.833}} \quad °C$$

$$P_{tot} = 8.96 + 12.71 = 21.67 \ W$$

$$\Delta T = (21670 / 202.2)^{0.833} = 49 \ °C$$

Efforts should be made to make $P_{core} \approx P_{cu}$ to optimize the design.

Note: P_{tot} in mW

Step #16 Determine the effect of the dc Bias on the value of inductance.

The procedure so far has assumed that there will be no appreciable drop in inductance factor A_L with increasing magnetizing force, H (Amps-Turns). However, as the magnetizing force increases, the inductance will decrease as the core moves further into saturation.

For the particular core and gap size chosen, we can check the value of inductance under peak magnetizing condition from the characteristic graphs in the POWERLITE C-Core Technical Bulletin, modifying the design where necessary.

Referring to the graph of A_L vs. H in the POWERLITE C-Core Technical Bulletin for the **AMCC-25**, we see that for the above design

$$L(0 \ Amp) = 375 \ \mu H$$

$$L(38.5 \ Amp) = 320 \ \mu H \ (85\% \ L_{INIT})$$

The relationship between the area product. W_aA_c of a core and its capability for storing energy can be obtained by using Faraday's Law of Induction.

$$V_L = N \frac{d\Phi}{dt} \quad \text{Volts} \tag{A.1}$$

$$N A_c' \frac{dB}{dt} \quad \text{Volts} \tag{A.2}$$

$$\text{where } \Phi = B A_c' \quad \text{Webers} \tag{A.3}$$

The voltage may be time-dependent, thus equation A.1 can be expressed as follows:

$$V_L = L \frac{di}{dt} \quad \text{Volts} \tag{A.4}$$

Combining equations A.2 and A.4

$$V_L = L \frac{di}{dt} = N A_c' \frac{dB}{dt} \quad \text{Volts} \tag{A.5}$$

Solving for the inductance L in equation A.5

$$L = N A_c' \frac{dB}{di} \quad \text{Henry} \tag{A.6}$$

Ampere's Law describes the relationship between electric current and magnetic field.

$$\int H \, d\ell = N I \quad \text{amperes-turns} \tag{A.7}$$

Solving the magnetic field intensity, H , in equation A.7

$$H = \frac{N I}{\ell'} \quad \text{amperes-turns/meter} \tag{A.8}$$

Solving for the current in equation A.8

$$I = \frac{H \ell'}{N} \quad \text{Amperes} \tag{A.9}$$

From the magnetic characteristic

$$B = \mu H \quad \text{Tesla} \tag{A.10}$$

The current of the wound coil can be defined as:

And, substituting equation A.8 into A.10

$$B = \frac{\mu N I}{\ell'} \quad \text{amperes-turns/meter} \tag{A.11}$$

Solving for μ and I in equation A.11

$$\mu = \frac{B \ell'}{N I} \quad \text{Henry/meter} \tag{A.12}$$

And

$$I = \frac{B \ell'}{\mu N} \quad \text{Amperes} \tag{A.13}$$

Substituting equation A.11 into equation A.6, we can determine the value of L in magnetic terms.

$$L = \frac{N^2 A_c' \mu}{\ell} \quad \text{Henry} \tag{A.14}$$

To define the relationship between the electrical energy the magnetic energy, we use the electrical Law:

$$P = VI \quad \text{Watts} \tag{A.15}$$

Substituting equation A.4 into equation A.15

$$P = V L I = L \frac{di}{dt} I \quad \text{Watts} \tag{A.16}$$

And solving for the energy W

$$W = \int P \, dt = \int L I T \frac{di}{dt} dt \quad \text{Watt-seconds}$$

$$W = 0.5 \; L \; I^2 \quad \text{Joule} \tag{A.17}$$

Substituting the value of L , determined into equation A.14 into equation A.17 , we get

$$W = \frac{\mu \; A_c' \; N^2 I^2}{2\ell'} \quad \text{Watt – seconds} \tag{A.18}$$

$$I = \frac{K W_a' J'}{N} \quad \text{Amperes} \qquad (A.19)$$

Equating A.19 an A.13

$$\frac{K W_a' J}{N} = \frac{B\ell'}{\mu N} \qquad (A.20)$$

And solving for μ,

$$M = \frac{B\ell'}{K W_a' J'} \quad \text{Henry/meter} \qquad (A.21)$$

Substituting the value of the permeability, μ, in equation A.21 and the current, I , of equation A.19 into the energy equation A.18 we arrive at:

$$W = \frac{B_m A_c' W_a' K J'}{2} \quad \text{Watt-seconds} \qquad (A.22)$$

And solving for $W_a'A_c'$

$$W_a'A_c' = \frac{2W}{B_m KJ'} \quad m^4 \qquad (A.23)$$

Where:

V = Voltage (V)
L = Inductance (H)
Φ = Flux (Wb)
B_m = Flux Density (T)
I = Current (A)
A_c' = Cross-section (m^2)
W_a' = Winding Area (m^2)
J' = Current Density (A/ $m^{2)}$
K = Window Fill Factor
W' = Energy (watt-seconds)
ℓ_m' = Magnetic Path Length(m)
μ = Permeability (H/m)

For $W_a (cm^2) = W_{a'} (m^2) \times 10^4$
$A_c(cm^2) = Ac (m^2) \times 10^4$
$J(A/cm^2) = J' (A/m^2) \times 10^{-4}$
$\ell m (cm) = \ell m'(m) \times 10^2$

$$W = \frac{B_m A_c W_a KJ \times 10^{-4}}{2}$$

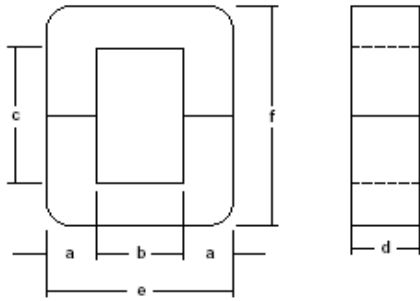Finally,

$$\frac{W_a A_c - 2 W (10^4)}{B_m KJ} \quad cm4$$

Ordering Information

Example:

AMCC-1000

Amorphous Metal ——
C-Core ————— ———— Part Designation

C-Cores Standard Core Sizes

CORE NUMBER	a (mm)	b (mm)	c (mm)	d (mm)	e (mm)	f (mm)	l_m (cm)	A_c (cm²)	MASS (g)	VOL (cm³)	W_a (cm²)	W_aA_c (cm⁴)
AMCC-6.3	10 ± 0.5	11	33	20 +0.5	31 +1	53 + 2	13.1	1.59	150	20.9	3.6	5.8
AMCC-8	11 ± 0.8	13	30	20 + 0.5	35 +1	52 + 2	13.2	1.79	170	23.7	3.9	7.0
AMCC-10	11 ± 0.8	13	40	20 + 0.5	35 +1	62 + 2	15.4	1.81	200	27.9	5.2	9.4
AMCC-16A	11 ± 0.8	13	40	25 + 0.5	35 +1	62 + 2	15.1	2.31	250	34.8	5.2	12.0
AMCC-16B	11 ± 0.8	13	50	25 + 0.5	35 +1	72 + 2	16.9	2.31	280	38.0	6.5	15.0
AMCC-20	11 ± 0.8	13	50	30 + 0.5	35 +1	72 + 2	17.5	2.71	340	47.4	6.5	17.6
AMCC-25	13 ± 0.8	15	56	25 + 0.5	41 +1	82 + 2	19.6	2.70	380	52.9	8.4	22.7
AMCC-32	13 ± 0.8	15	56	30 + 0.5	41 +1	82 + 2	20	3.20	460	64.1	8.4	26.9
AMCC-40	13 ± 0.8	15	56	35 + 0.5	41 +1	82 + 2	19.9	3.71	530	73.8	8.4	31.2
AMCC-50	16 ± 1	20	70	25 + 0.5	52 +1	102 + 3	24.9	3.30	590	82.2	14.0	46.2
AMCC-63	16 ± 1	20	70	30 + 0.5	52 +1	102 + 3	25.3	3.91	710	98.9	14.0	54.7
AMCC-80	16 ± 1	20	70	40 + 1	52 +1	102 + 3	25.4	5.21	950	132	14.0	72.9
AMCC-100	16 ± 1	20	70	45 + 1	52 +1	102 + 3	25	5.91	1060	148	14.0	82.7
AMCC-125	19 ± 1	25	83	35 + 1	63 +1	121 + 3	30.2	5.40	1170	163	20.8	112
AMCC-160	19 ± 1	25	83	40 + 1	63 +1	121 + 3	28.5	6.50	1330	185	20.8	135
AMCC-200	19 ± 1	25	83	50 + 1	63 +1	121 + 3	29.8	7.81	1670	233	20.8	162
AMCC-250	19 ± 1	25	90	60 + 1	63 +1	128 + 3	31.4	9.31	2100	292	22.5	210
AMCC-320	22 ± 1	35	85	50 + 1	79 +1	129 + 4	32.5	9.30	2170	302	29.8	277
AMCC-400	22 + 1	35	85	65 + 1	79 +1	129 + 4	33.6	11.7	2820	393	29.8	348
AMCC-500	25 ± 1	40	85	55 + 1	90 +1	135 + 4	35.6	11.3	2900	404	34.0	386
AMCC-630	25 ± 1	40	85	70 + 1	90 + 1	135 + 4	35.6	14.4	3670	511	34.0	488
AMCC-800A	25 ± 1	40	85	85 ± 1.5	90 + 1	135 + 4	35.6	17.4	4450	620	34.0	592
AMCC-800B	30 ± 1	40	95	85 ± 1.5	100 + 1	155 + 4	39.3	21.0	5930	826	38.0	799
AMCC-1000	33 ± 1	40	105	85 ± 1.5	106 + 1	171 + 5	42.7	23.0	7060	983	42.0	967

l_m = mean magnetic path length   Ac = net cross-sectional area