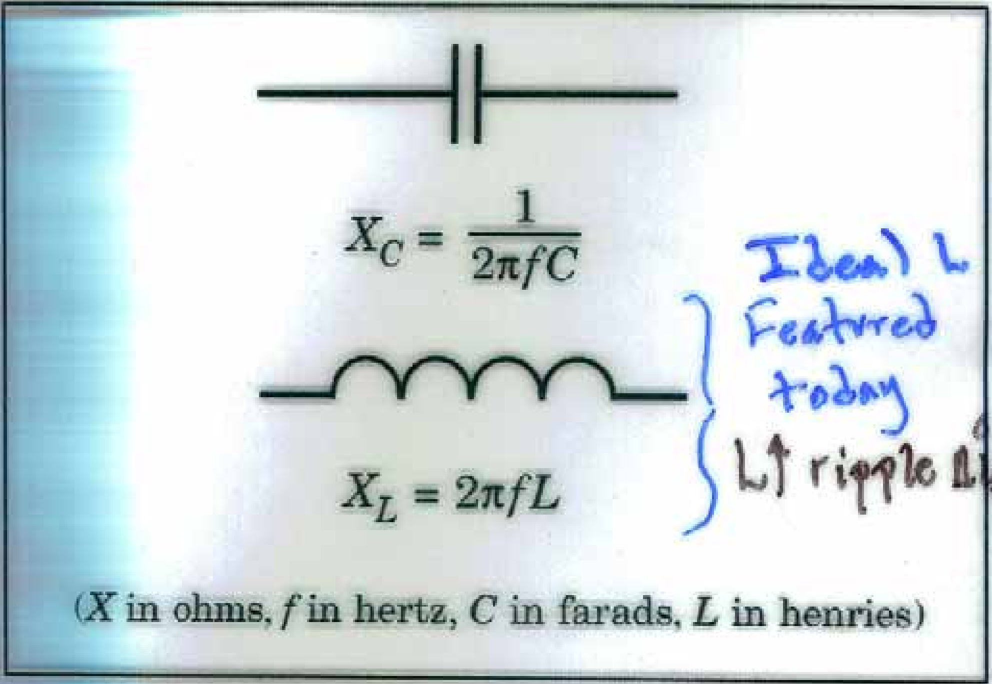
ECE 562

Week 4 Lecture 2

Fall 2008

Week 4 Lecture 2
Summary

- Section notes
  - Slides 3-8 - Inductor overview
  - Slides 9-17 - Inductor properties
  - Slides 18-37 – Commercial inductors
  - Slides 37-60 – Current and saturation
  - Slides 61-69 – Inductor ratings
  - Slides 69- 77– Inductor sizing
  - Slides 78-80 – Resistors
  - Slides 81-86 – In-circuit properties
  - Slides 87- 98 – Inductor role in circuits



$$X_C = \frac{1}{2\pi f C}$$

$$X_L = 2\pi f L$$

(X in ohms, f in hertz, C in farads, L in henries)

Ideal L
Featured
today

L↑ ripple ΔI↓

Figure 1 · The simplest frequency-dependent behaviors are capacitive and inductive reactance.

Specs: I_{max} (sat), DCR / Other?

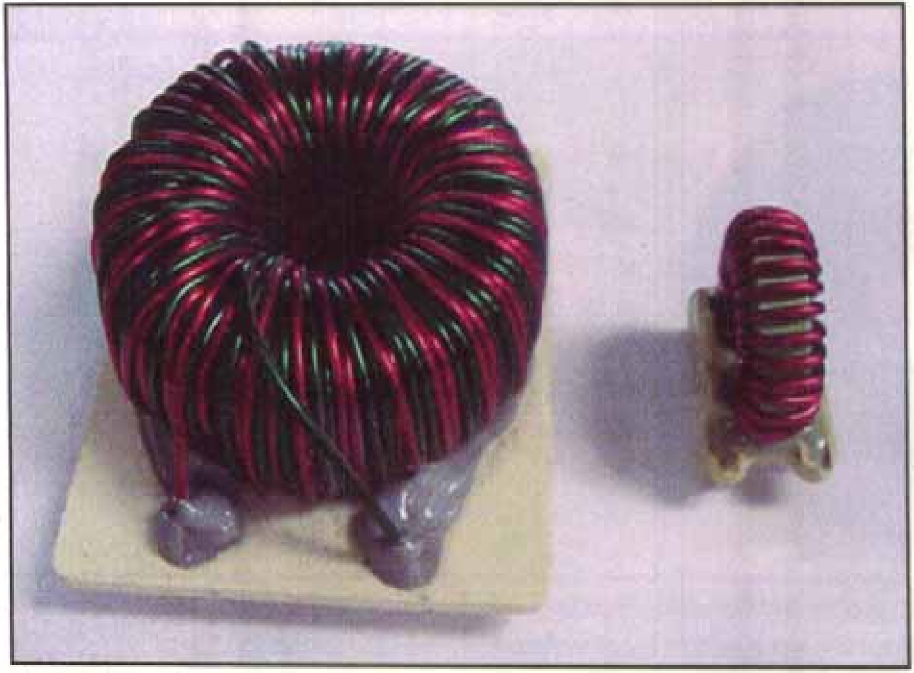


Fig. 3. *The multiple-layer inductor (left) has so much parasitic capacitance that an additional inductor is needed as a filter (right) to eliminate noise that passes through the multiple-layer inductor. Together, both inductors consume a large amount of board space and add considerable size, weight and cost to the design.*

ler obviously reduces e

Actual R model $f \to$ MHz

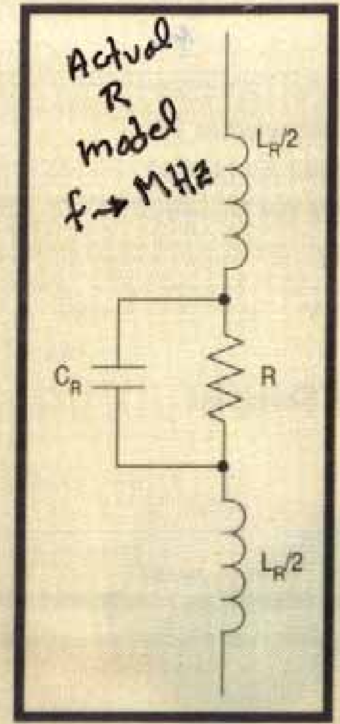
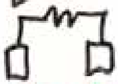


Fig. 1. *Actual resistors have parasitic capacitance and inductance.*

$\boxed{L}$

- wire wound R have high "L"
- R leads add L
- surface mount R lower L

$\boxed{C}$



c of leads

- surface mount C ↑

f_R exists for each L

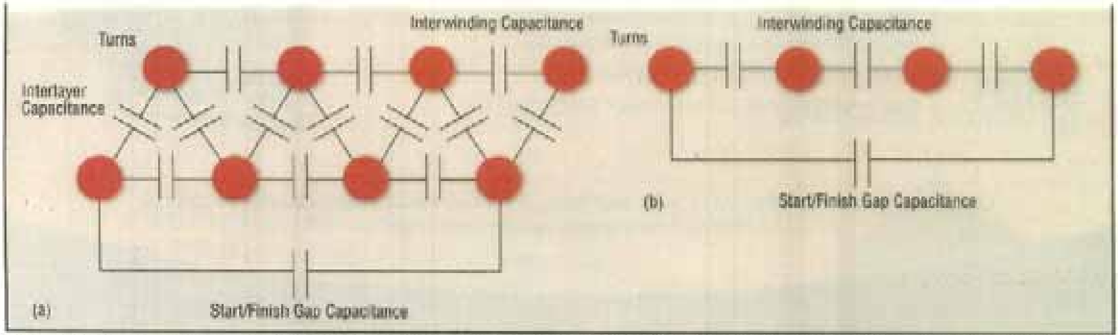
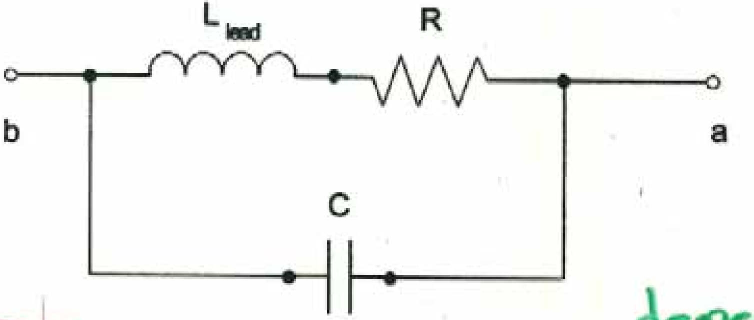


Fig. 1. *An inductor having multiple layers of windings (a) exhibits much more parasitic capacitance than one with just a single winding layer (b).*

①Depends on How made ②/100 effects

f_{SR} is self resonant frequency



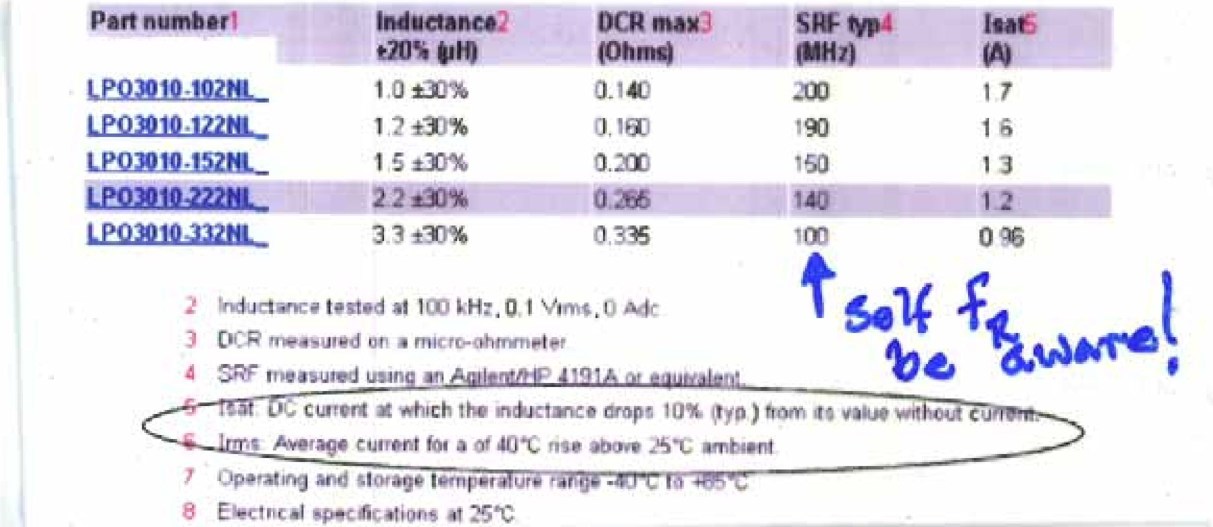
depends on ?

VS

depends on geometory materials

Figure 18. Simplified equivalent circuit for inductor.

L(PCB) ≠ L(Core)



Part number[1]	Inductance[2] ±20% (µH)	DCR max[3] (Ohms)	SRF typ[4] (MHz)	Isat[5] (A)
LPO3010-102NL	1.0 ±30%	0.140	200	1.7
LPO3010-122NL	1.2 ±30%	0.160	190	1.6
LPO3010-152NL	1.5 ±30%	0.200	150	1.3
LPO3010-222NL	2.2 ±30%	0.266	140	1.2
LPO3010-332NL	3.3 ±30%	0.335	100	0.96

↑ self f_R
be aware!

2 Inductance tested at 100 kHz, 0.1 Vrms, 0 Adc
3 DCR measured on a micro-ohmmeter
4 SRF measured using an Agilent/HP 4191A or equivalent.
5 Isat: DC current at which the inductance drops 10% (typ.) from its value without current.
6 Irms: Average current for a of 40°C rise above 25°C ambient
7 Operating and storage temperature range -40°C to +85°C
8 Electrical specifications at 25°C

DCR is R(wire) @ DC
Higher f same R?

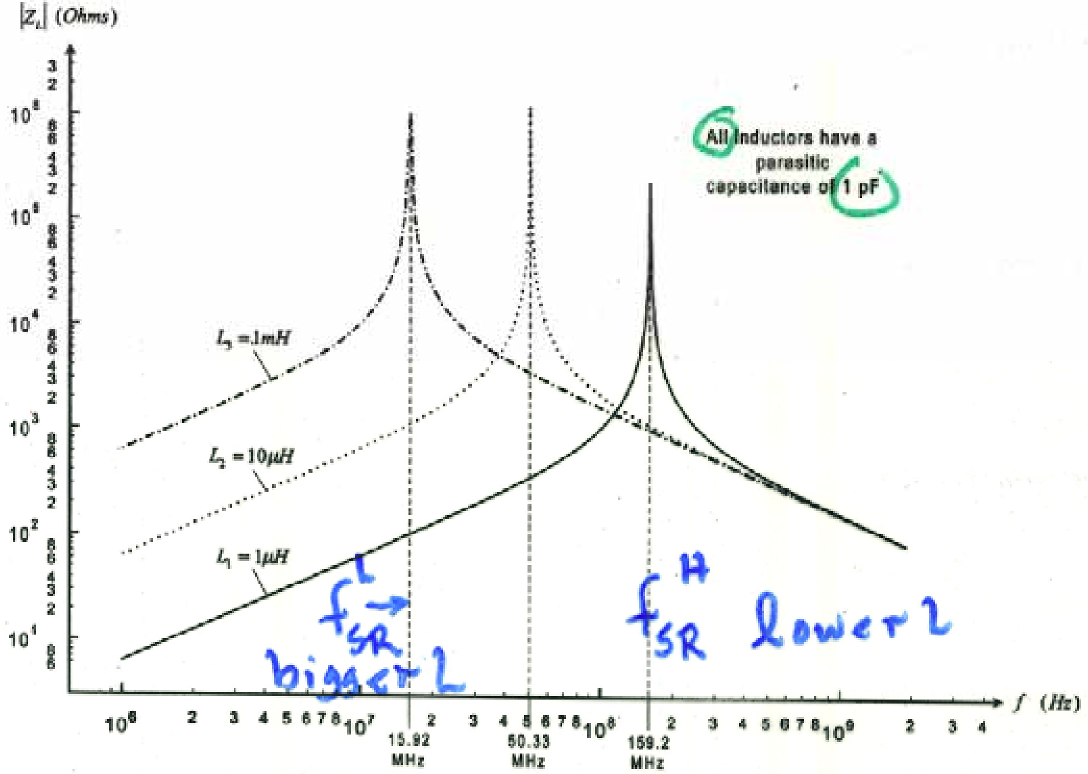


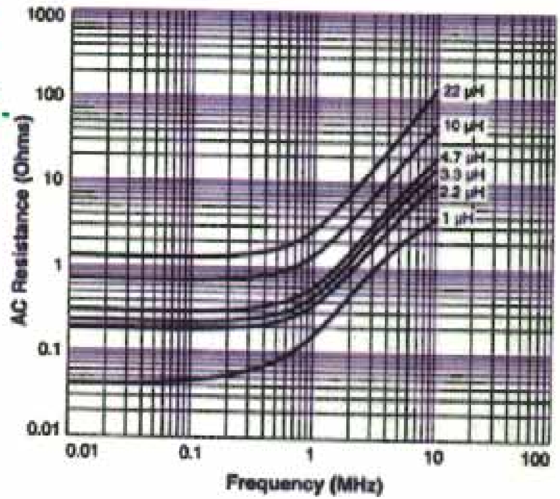
Figure 19. Plot of frequency dependent behavior of equivalent circuits for various inductors.

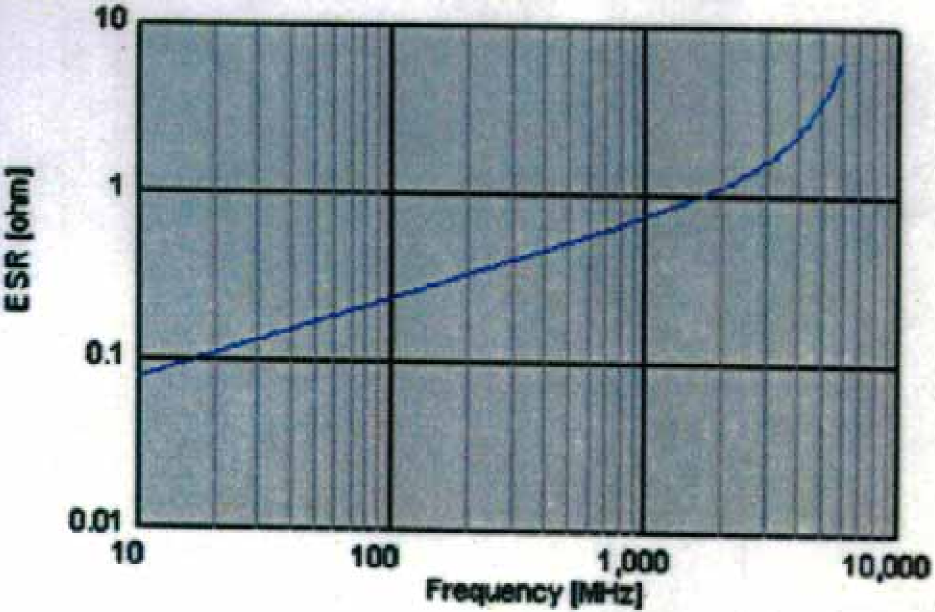
- Inherent to any conducting wire is an element of resistance. Also, the insulating material added around the wire, to prevent short circuit from one turn of the coil to another, acts as a dielectric that adds capacitance between the turns. So, each inductor comes with an inherent resistance and capacitance. In addition, at higher frequencies, current tends to flow closer to the conductor surface, an effect known as the skin effect. Thus, the total impedance offered by an inductor varies with voltage and frequency applied to it. An example inductor impedance vs frequency graph is shown below.

R(inductor wire) vs f

wire insulation → ε

needed to max. N^2


Typical AC Resistance vs Frequency



(c)

Definitions

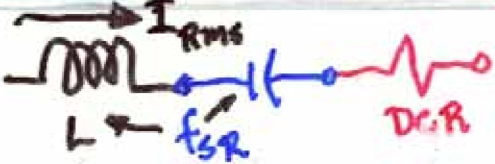
L – Inductance The primary functional parameter of an inductor. This is the value that is calculated by converter design equations to determine the inductors ability to handle the desired output power and control ripple current.

DCR – DC Resistance The resistance in a component due to the length and diameter of the winding wire used.

SRF – Self Resonant Frequency The frequency at which the inductance of an inductor winding resonates naturally with the distributed capacitance characteristic of that winding. $W = 1/\sqrt{LC}$

Isat – Saturation Current The amount of current flowing through an inductor that causes the inductance to drop due to core saturation.

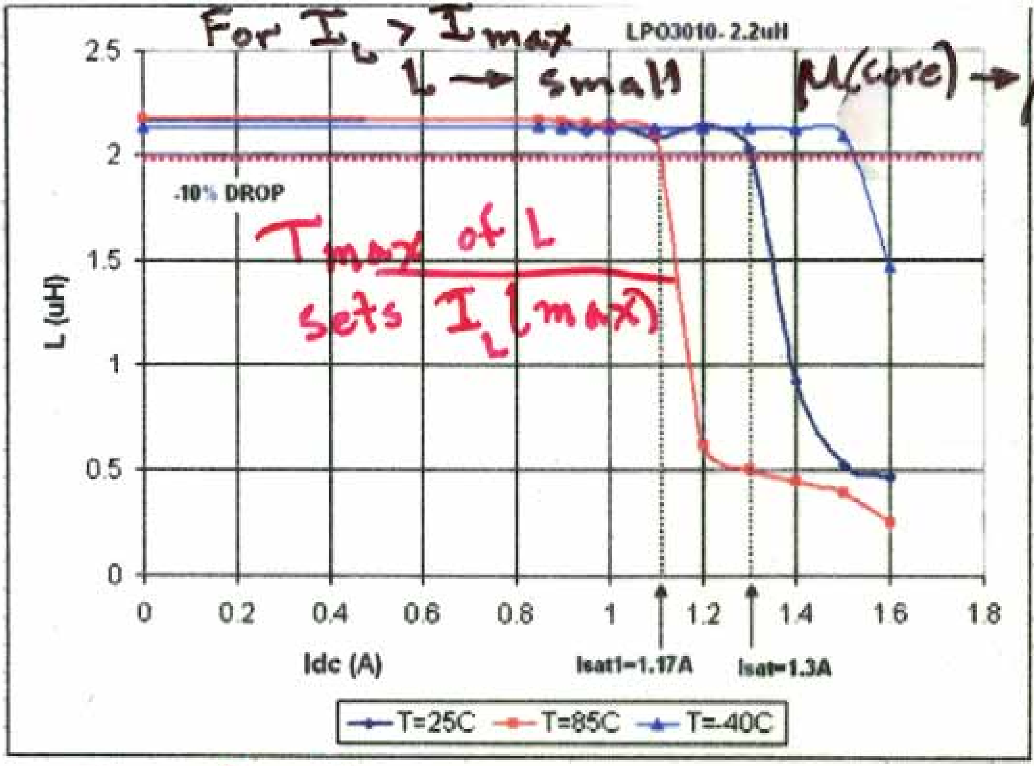
Irms – RMS Current The amount of continuous ~~current flowing~~ through an inductor that causes the maximum allowable temperature rise.



For $I_L > I_{max}$

$L \to$ small

$\mu(\text{core}) \to \mu_0$

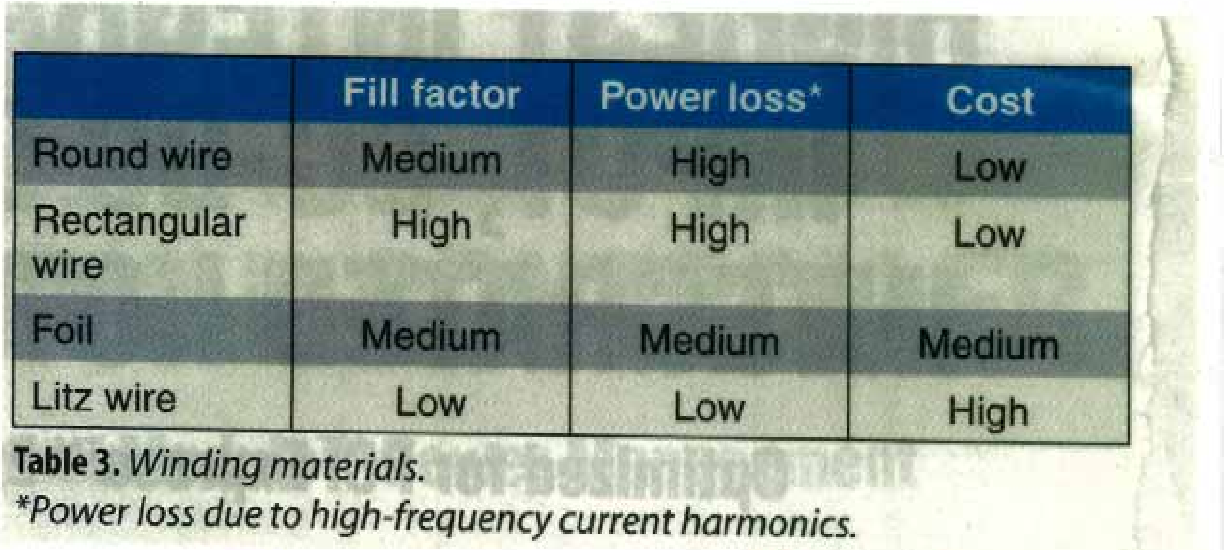


T_{max} of L sets $I_L(max)$

T_{max} for L

Heat in: Core loss / $I^2 R_{wire}$

Heat out: Heatsinks



	Fill factor	Power loss*	Cost
Round wire	Medium	High	Low
Rectangular wire	High	High	Low
Foil	Medium	Medium	Medium
Litz wire	Low	Low	High

Table 3. *Winding materials.*

**Power loss due to high-frequency current harmonics.*

Proximity Effects Φ_ℓ and wires

$$R_{AC}(\text{wire}) \simeq \begin{matrix}10\\100\end{matrix} R_{DC}(\text{wire})$$

The Current Ratings of an Inductor

We know that the "energy handling capability" of an inductor, 1/2 x L x Ipeak^2 , is one way of picking the size of the inductor. But most vendors do not provide this number upfront.

However, they do provide one or more "current ratings". And if we interpret these current rating(s) correctly, that serves the purpose too.

The current rating may be expressed by the vendor either as a maximum rated

- I_{DC}

or a maximum rated

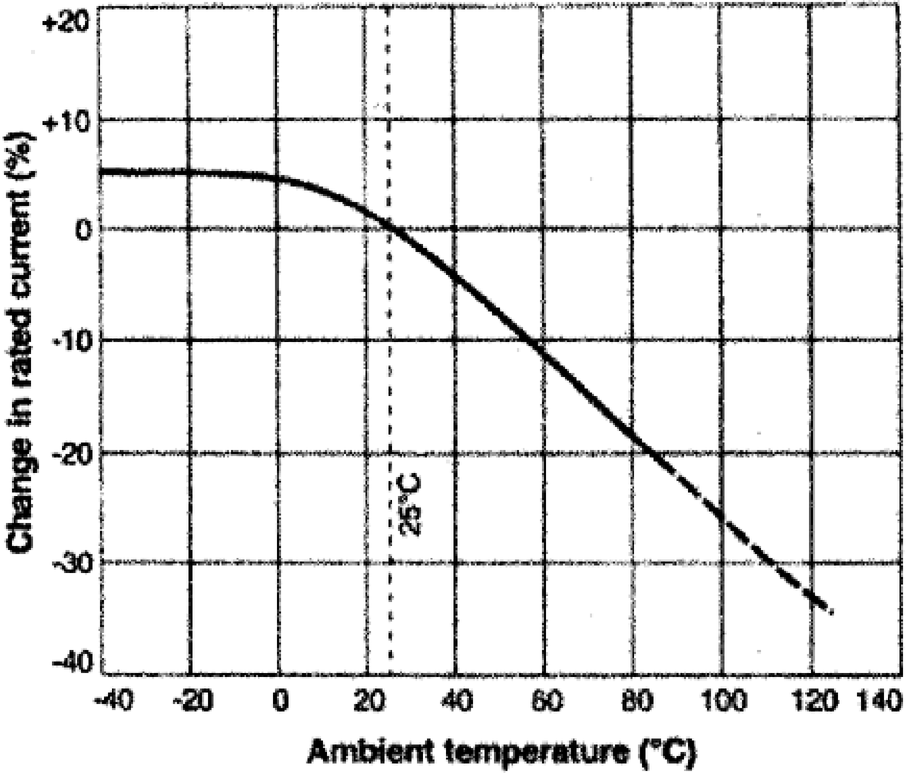
- I_{RMS}

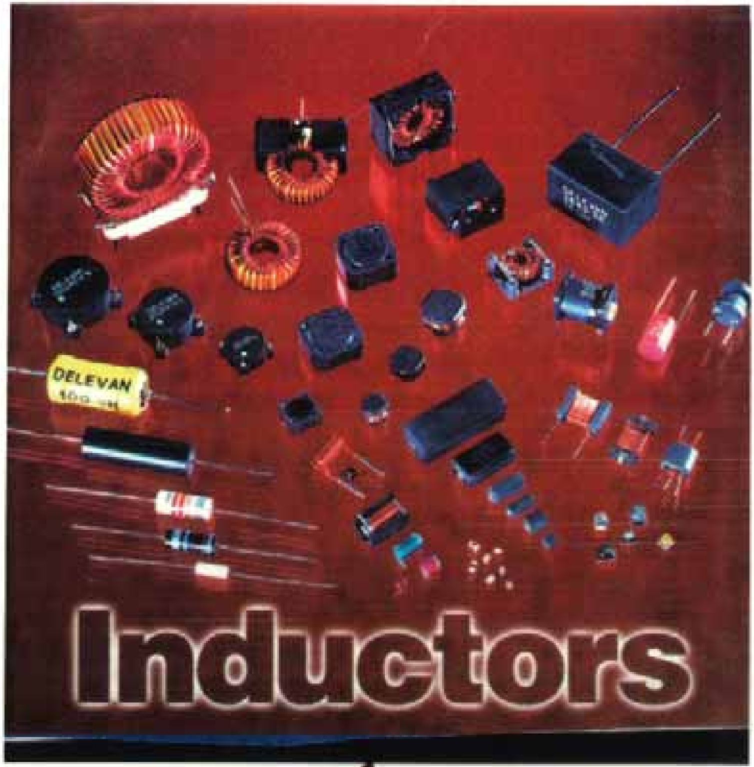
or/and a maximum saturation current

- I_{SAT}

The first two are usually considered synonymous, since *the RMS and dc values of a typical inductor current waveform are almost equal* (more some when we have continuous conduction mode combined with typically low inductor current ripple).

Current Derating



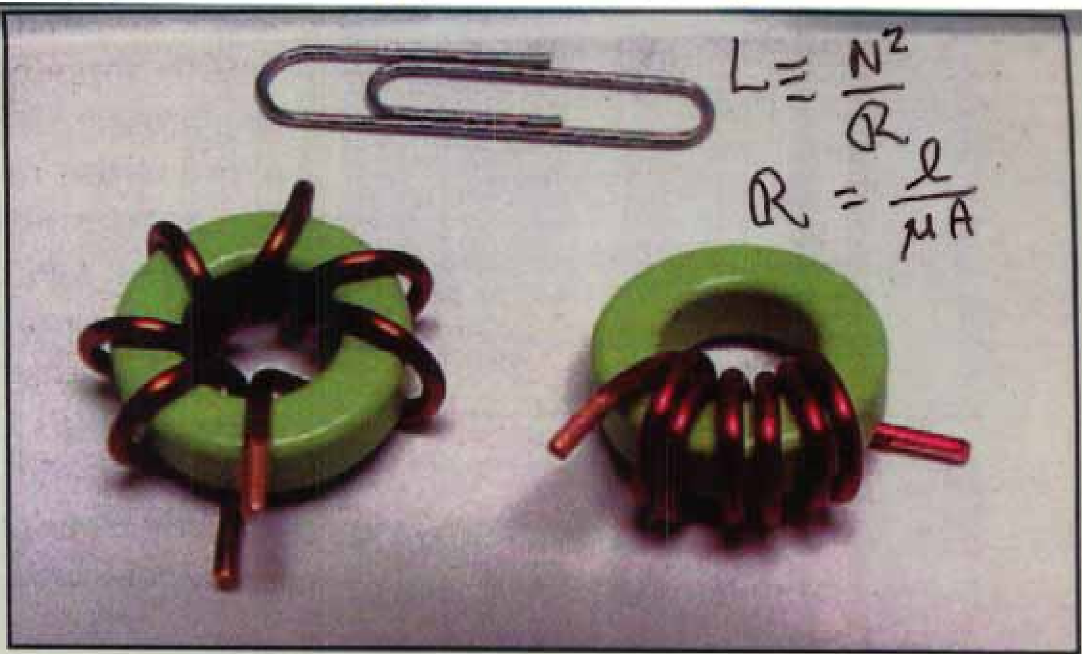


$$L = \frac{N^2 (\text{wire})}{\mathcal{R}_{core}}$$

← Cu (next to $N^2 (\text{wire})$)

← iron ferrite core gaped? (next to $\mathcal{R}_{core}$)

Both 6 Turns

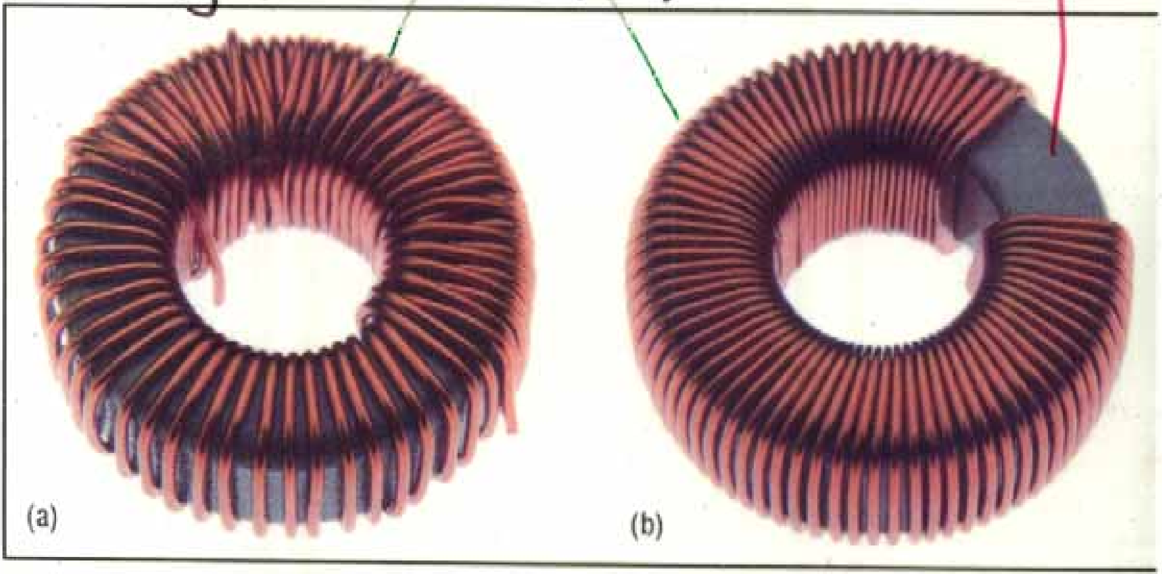


$$L \cong \frac{N^2}{\mathcal{R}}$$

$$\mathcal{R} = \frac{l}{\mu A}$$

Fig. 6. *Toroid inductors can vary in inductance values based on how they are wound. The inductor on the left has a more uniform winding distribution of inductance, while the one on the right has a higher inductance value.*

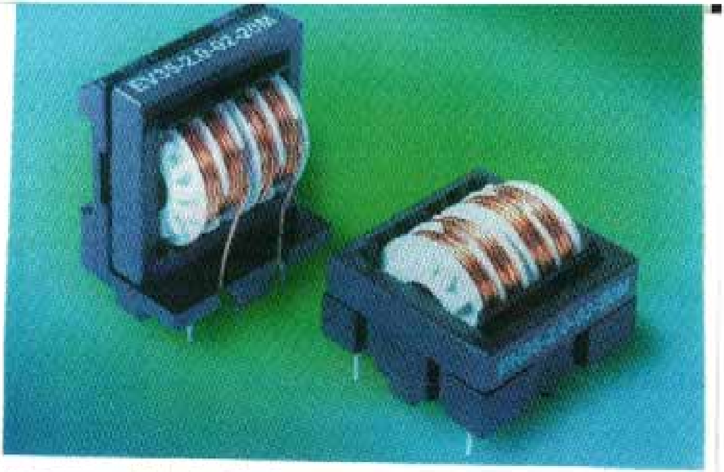
Big N $\Rightarrow$ $L \sim N^2 \uparrow$



Cu

Ferrite Core

Fig. 2. *A multiple-layer toroid with no start/finish gap (a) has significantly more parasitic capacitance than a single-layer toroid with a start/finish gap (b).*

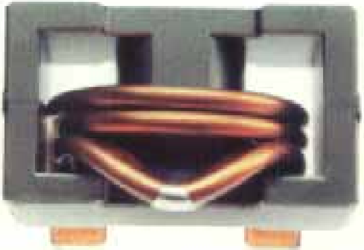


Common-Mode Chokes

Compact common-mode chokes are designed to provide high inductance for suppression of electromagnetic interference. The EV and EH chokes are available in inductance values up to 90 mH and in currents up to 5 A. Two-section winding allows high-frequency resonance, and broadband attenuation characteristics provide a stable frequency response. The RoHS-compliant chokes have a flammability rating of UL 94V-0. At 50°C, the chokes have a maximum operating voltage of 250 V. Power operating frequency ranges from dc to 400 Hz. The EV chokes offer a narrow width, and the EH chokes feature a low profile. **Schaffner EMC, Edison, NJ, 732/225-9533.**

www.schaffnerusa.com

Our new 100 Amp power inductors should make everyone smile

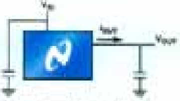


High I A (core)

Big L via μ (core)

$$L = \frac{N^2}{\mathcal{R}} = \frac{N^2 \mu_c A_{core}}{l_{core}}$$

Linear regulator



Function: Step-down ($V_{OUT} < V_{IN}$)

When to use: Typically when $I_{OUT} < 1A$, ultra low-dropout, and low-noise applications

Characteristics: Excellent option where fixed output, low current, and low voltage drops are required. Easy to implement

Devices to use: Any low dropout, linear regulator

Comments: Great for micropower applications

Non-synchronous buck

Function: Step-down ($V_{OUT} < V_{IN}$)

When to use: Typically when V_{IN} is 3x to 5x V_{OUT} and I_{OUT} is > 0.5A and < 5A

Characteristics: Easy to design and good efficiency for the above mentioned typical $V_{IN}/V_{OUT}/I_{OUT}$ conditions

Devices to use: All buck integrated regulators and controllers

Synchronous buck

Function: Step-down ($V_{OUT} < V_{IN}$)

When to use: When high efficiency is required with high-output current (> 5A) or low duty cycles (V_{IN} > 5 x V_{OUT} and/or I_{OUT} < 0.5A)

Characteristics: A second switch replaces the diode in the basic buck topology, reducing losses in the conditions mentioned above

Devices to use: Any "synchronous rectification" buck integrated regulator or controller

Figure 2: Step-down configurations

Small L in size
Uses PC Board for N

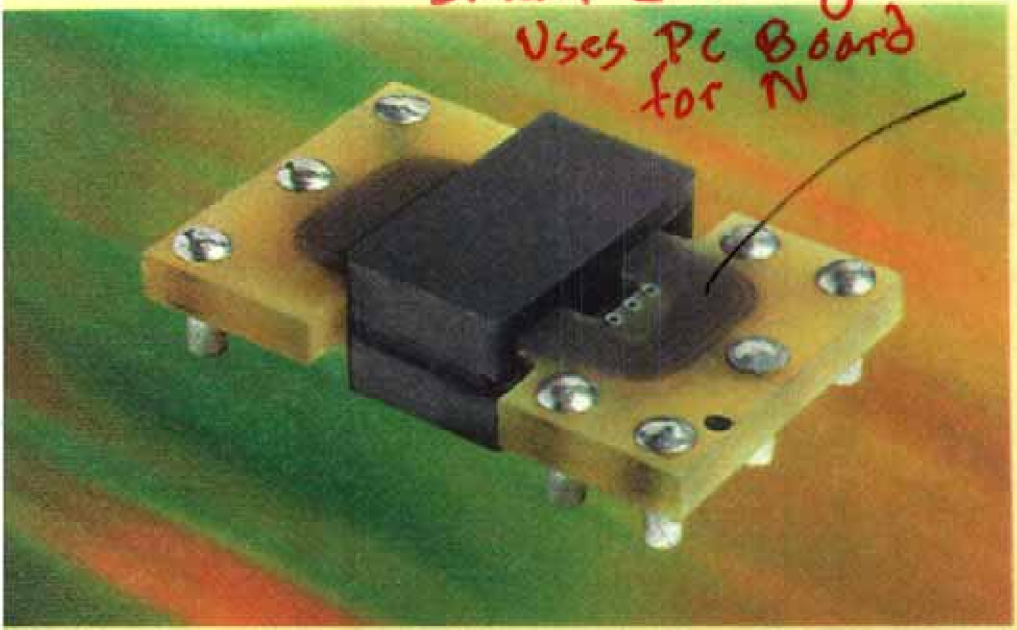


Fig. 3. *PCB-based planar inductors use magnetic cores mounted on the top and bottom of the board, and in many cases, run through it, as shown in this photo.*

Wire is Cu PCB traces

Cores are planar and inserted around PCB

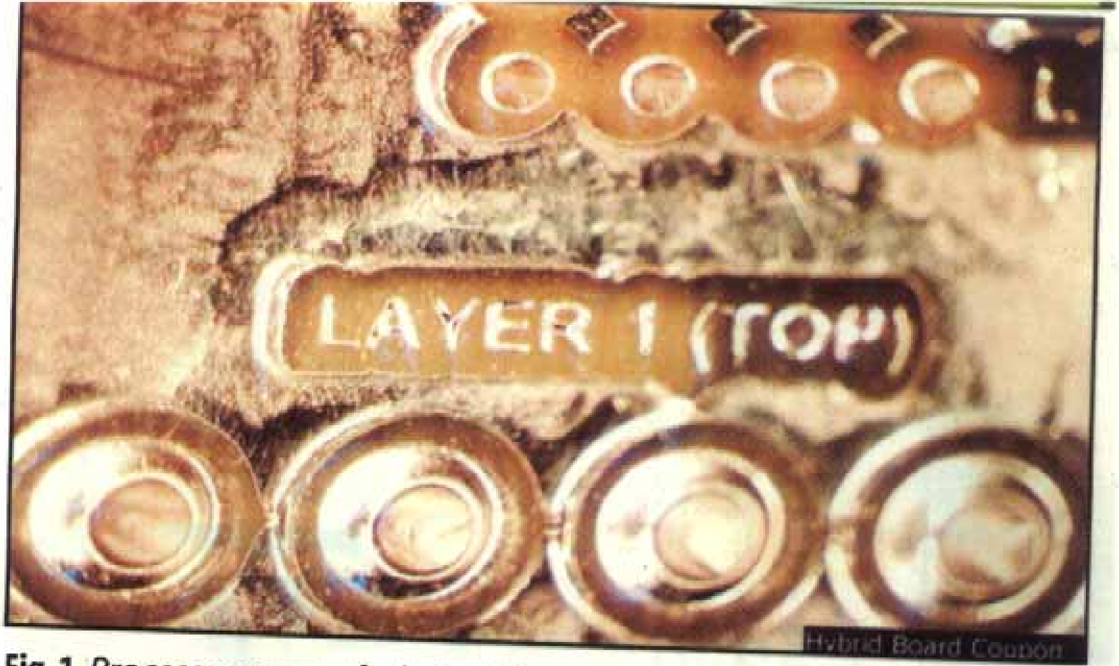


Fig. 1. *Process coupon of a hybrid thick/thin copper printed wiring board (solder coated).*

ΔT(°C)	Trace Width (mil)	Trace Thickness (oz/ft^2)	Location (external/internal)	Max Current (Amps)
20	50	1	External	2.629
20	50	10	External	9.137
20	500	1	External	14.663
20	500	10	External	50.958
20	2000	1	External	48.121
20	2000	10	External	167.237
50	50	1	External	3.996
50	50	10	External	13.888
50	500	1	External	22.288
50	500	10	External	77.459
50	2000	1	External	73.146
50	2000	10	External	254.209

Current carrying capacity of various copper trace sizes.

PCB
Copper

High f L

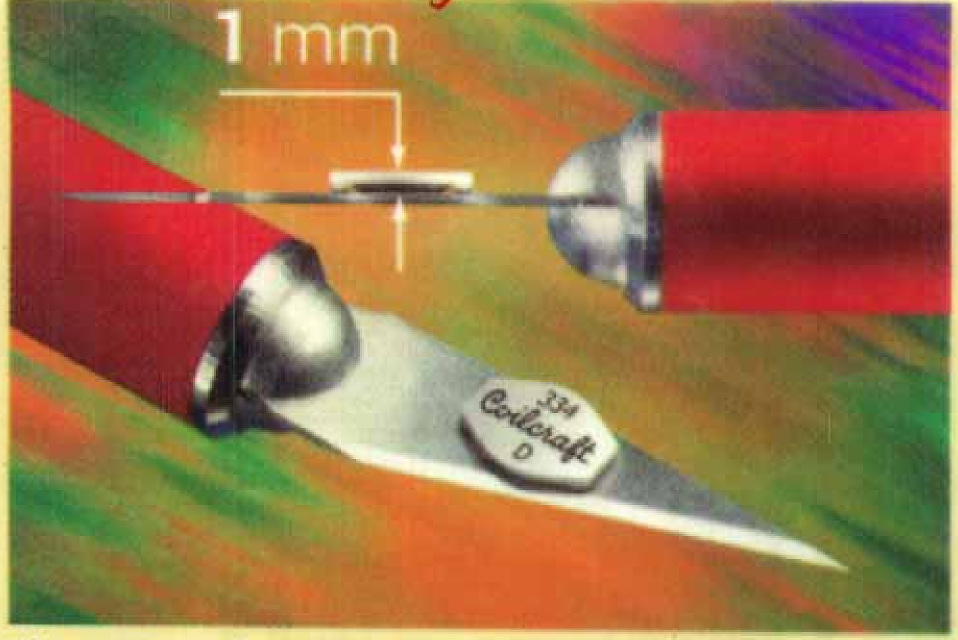
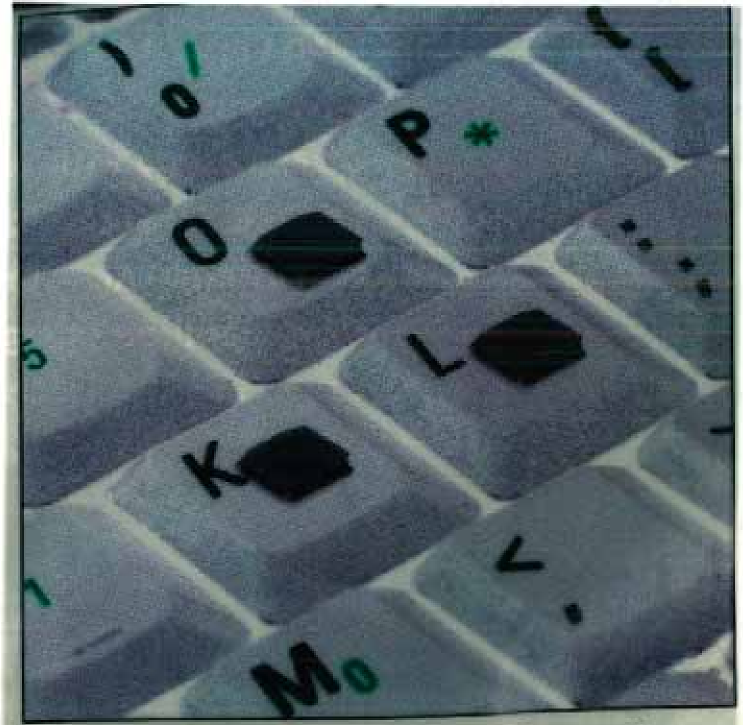


Fig. 1. *Transformers and inductors as thin as 1 mm are in demand.*

Driving Factors



Taiyo Yuden's NR6012 series of 6-mm x 6-mm x 1.2-mm power inductors offers current ratings as high as 1730 mA.

bussmann.com). Featuring

NEW

LPO1704 Power Wafer™

Inductance 1.2 - 330 μH
Isat up to 2.1 A, Irms up to 3.6 A
Kit C142 $60

DO1606 Power Wafer™

Inductance 1.0 - 1000 μH
Isat up to 2.5 A, Irms up to 2.3 A
Kit C138 $70

NEW

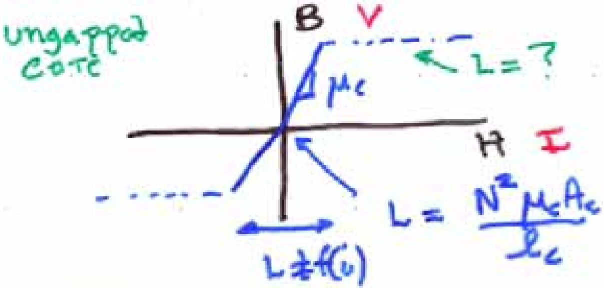
0805PS Series - Shielded

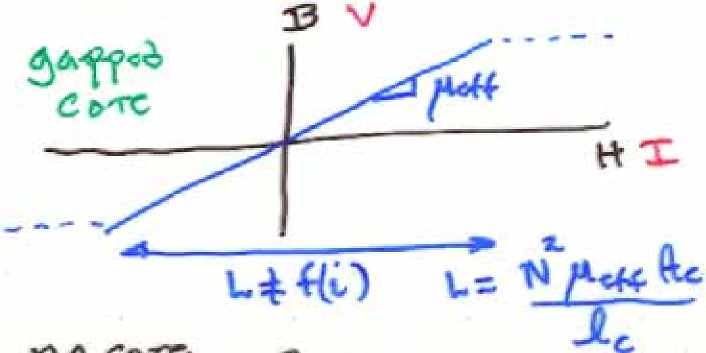
Inductance 1.0 - 330 μH
Isat up to .8 A, Irms up to 1.5 A
Kit C148 $80

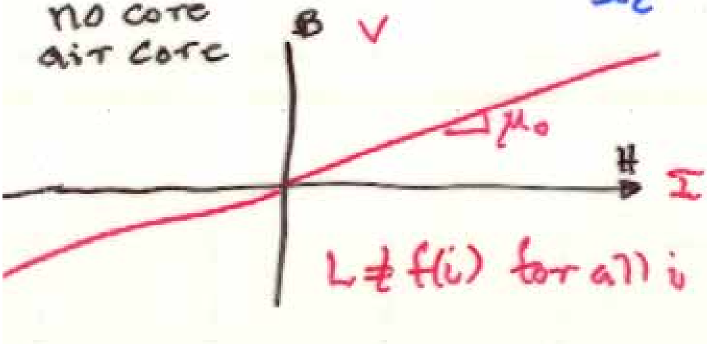
NEW

1008PS Series - Shielded

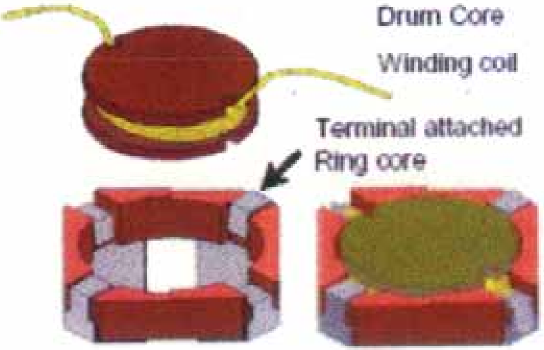
Inductance 1.0 - 1000 μH
Isat up to 3.0 A, Irms up to 2.0 A
Kit C141 $80


ungapped core
B
V
L = ?
μ_c
H
I
$L = \frac{N^2 \mu_c A_c}{l_c}$
$L \neq f(i)$


gapped core
B
V
μ_{eff}
H
I
$L \neq f(i)$
$L = \frac{N^2 \mu_{eff} A_c}{l_c}$


no core
air core
B
V
μ_0
H
I
$L \neq f(i)$ for all i

(1)Magnetic Shield



Drum Core

Winding coil

Terminal attached Ring core

(2)No Shield

Drum Core

Terminal

Winding coil

$? = N^2 \mu \left(\frac{A}{l}\right)$

EMI issues

Φ_e goes into Cu wires and forces skin effects

$R_{wire}(\Phi_e \neq 0) = 10 R_{wire}(\Phi=0)$

DS1608 Series - Shielded

Inductance 1.0 – 10,000 µH

Imax up to 3.0 A

Kit C115 $80

DS5022 Series - Shielded

Inductance 10 - 1000 µH

Isat up to 8 A, Irms up to 3.9 A

Kit C117 $60

Not fully shielded

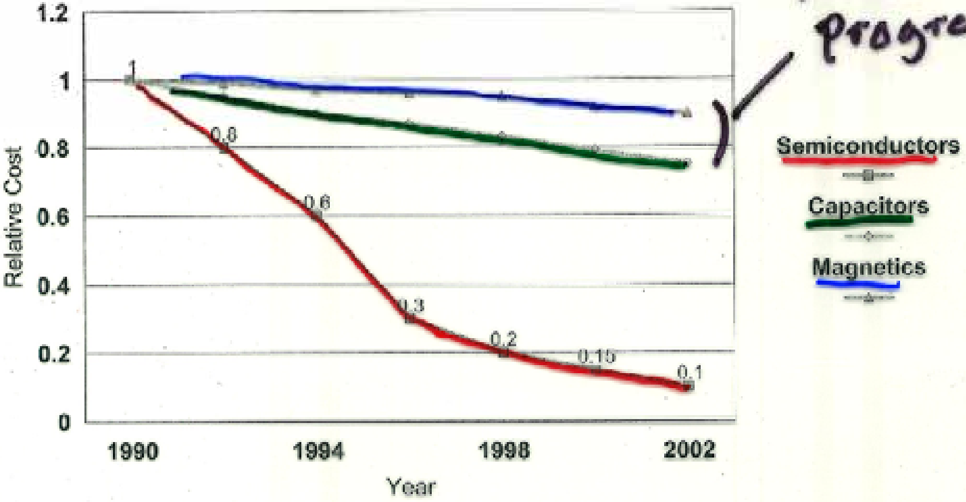
DO5022HC Series - High current

Inductance 0.78 - 15 µH

Isat up to 30 A, Irms up to 15 A

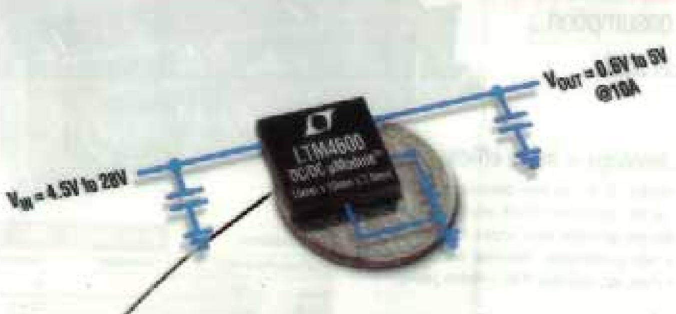
Kit C113 $60

Technology Cost Trend



Magnetics
Needs Progress!

Instant 10A Power Supply

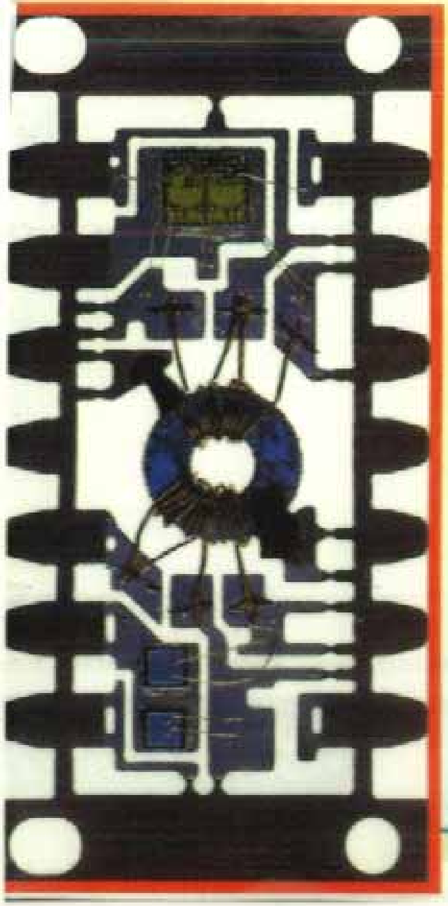


Complete, Quick & Ready.

Internal Q

L

But @ 100 A? **NO** IC



Prompted by advances in packaging technology, the trend for dc/dc converters leans toward enclosing more of the converter magnetics within an IC package, as shown in this example from TI/Burr-Brown.

5

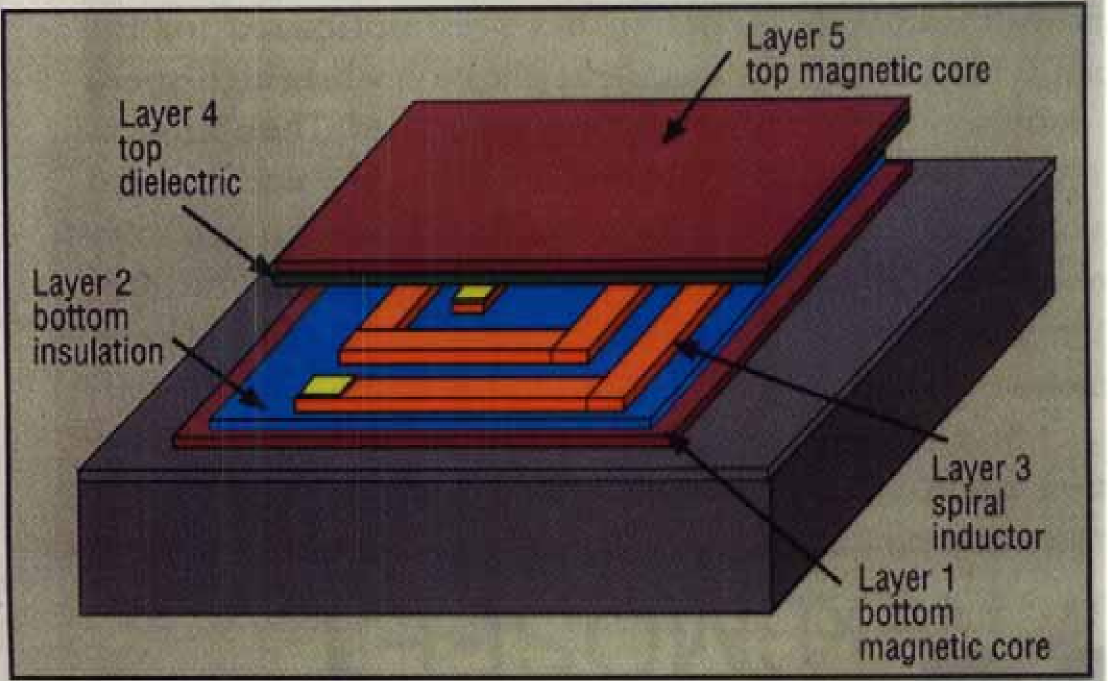


Fig. 3. *In Enpirion's MEMS-based inductor, a thick electroplated copper spiral coil is sandwiched between two planar magnetic layers.*

Our big L's in buck
boost
buck-boost

have "free wheeling i paths"

Other stray / trace L do not have free wheeling paths $\Longrightarrow$?

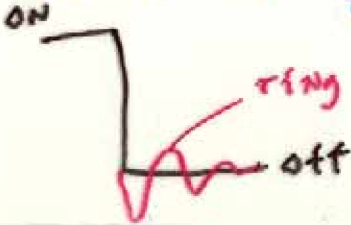
① $$V_{spikes} = \underbrace{L_{stray}}_{\text{Must } L\downarrow}(20nH)\ \underbrace{di/dt}_{\text{high during switch transitions}}\left(\frac{2A}{30ns}\right)$$

$\Delta V \approx 1.3\,V$

Killer? spike

via (wide / short) traces

② Ringing of switch voltage due to $f = \dfrac{1}{2\pi\sqrt{L_{stray}\,C(\text{device off})}}$



V_{max} cause device failure

The dc rating of an inductor is basically the dc current we can pass through it, such that we get a specified temperature rise (typically 40 to 55°C depending on the vendor).

The last rating, i.e. the I_{SAT} , is the maximum current we can pass, just before the core starts saturating.

limit stingy Avg $I < I_{SAT}$

- We will also find that many, if not most, vendors have chosen the wire gauge in such a manner that the I_{DC}/I_{RMS} and the I_{SAT} ratings of any inductor are also virtually the same. And therefore, they just publish one maximum (single) rating -- for example, "the inductor is rated for 5A".

I^2 DCR limit $< I_{SAT}$

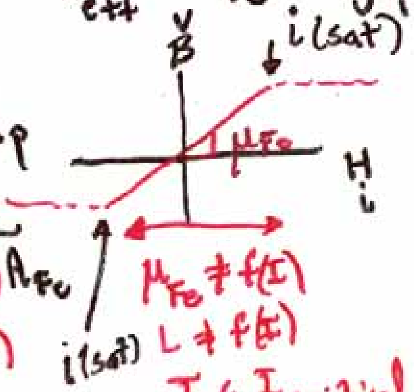
Basically, having determined the I_{SAT} of the inductor, the vendor has usually consciously tweaked the wire gauge (at this saturation current level), so as to also get the specified temperature rise. The rationale for doing so is as follows --- suppose the inductor had a dc rating of 3A and an I_{SAT} of 5A. The 5A rating is then likely to be *superfluous*, because users would probably never select this inductor for an application that required more than 3A anyway. Therefore, the excessive I_{SAT} rating in this case, essentially amounts to an *unnecessarily over-sized core*. Of course, if we do find an inductor with different I_{DC} and I_{SAT} ratings. it is possible the vendor may have

Stingy!

$$L = \frac{N^2}{\mathcal{R}_{eff}} \qquad \mathcal{R}_{eff} = \mathcal{R}_{Fe} + \mathcal{R}_{gap}$$

$$\mathcal{R}_{Fe} \gg \mathcal{R}_{gap}$$

$$\mathcal{R}_{Fe} = \frac{l_{Fe}}{\underbrace{\mu_{Fe}}_{f(I)} A_{Fe}}$$



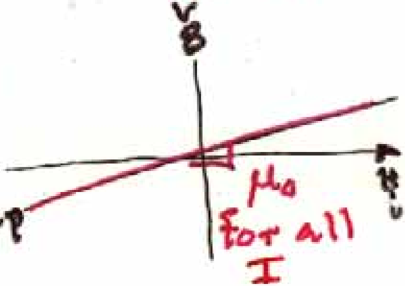
$\mu_{Fe} \neq f(I)$

$L \neq f(I)$

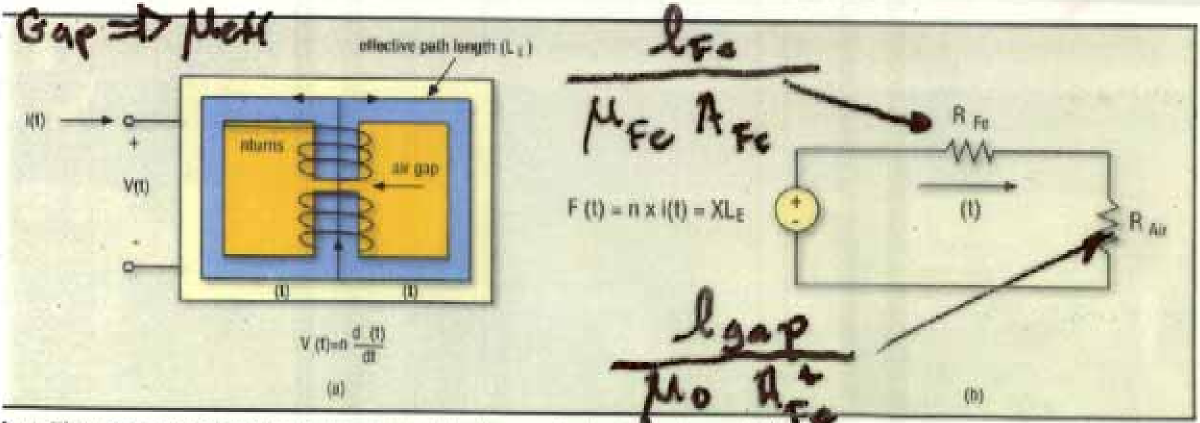
$I_L < I_{critical}$

$$\mathcal{R}_{gap} \gg \mathcal{R}_{Fe}$$

$$\mathcal{R}_g = \frac{l_g}{\underbrace{\mu_0}_{\neq f(I_L)} A_{gap}}$$



μ_0 for all I



Gap $\Rightarrow \mu_{eff}$

$$\frac{l_{Fe}}{\mu_{Fe} A_{Fe}}$$

$$\frac{l_{gap}}{\mu_0 A_{Fe}}$$

ig. 1. *This magnetic circuit (a) is represented by the equivalent circuit model (b).*

Air gaps
are distributed
throughout

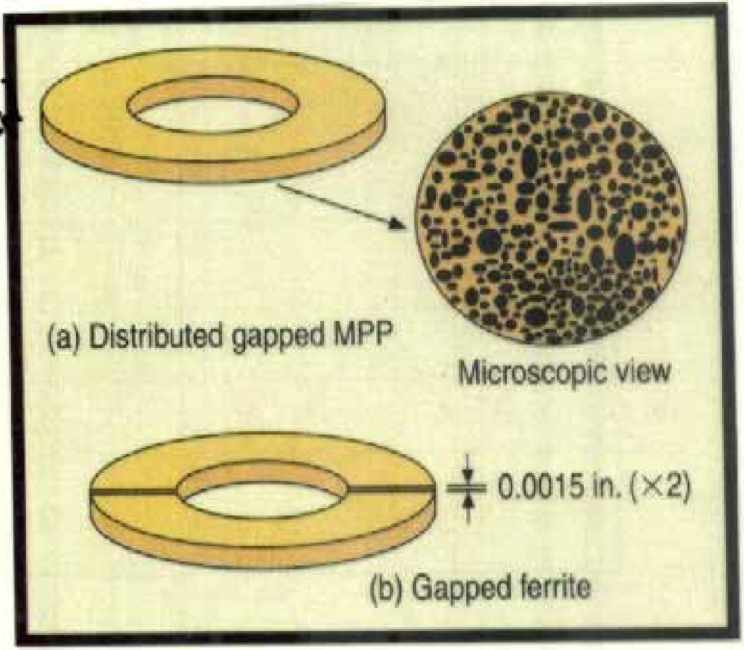
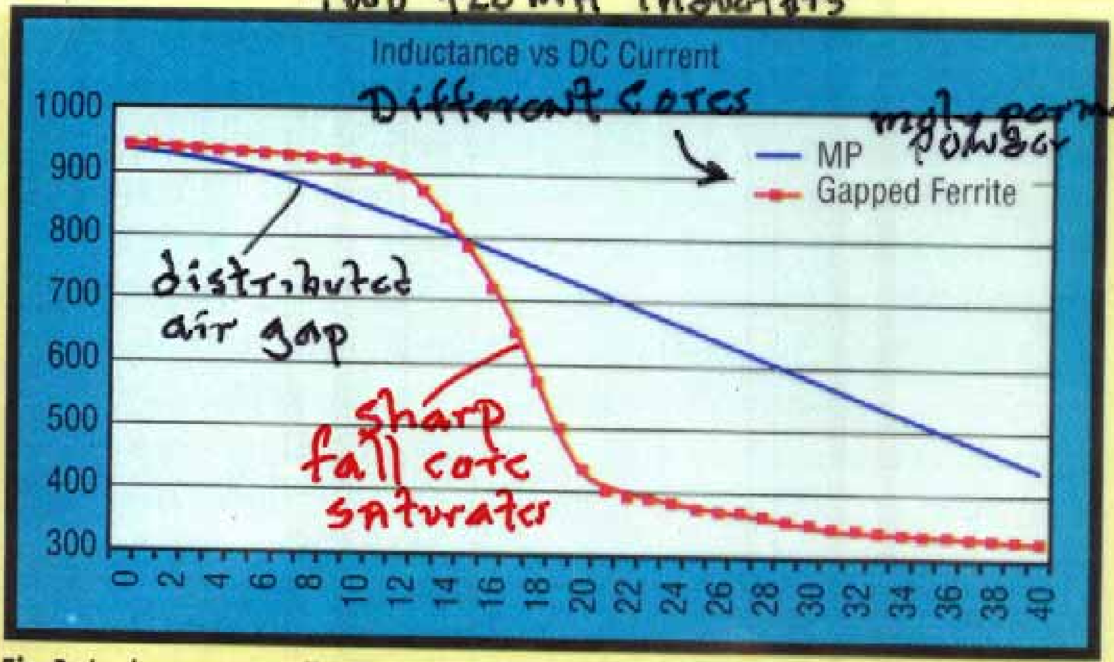


Fig. 3. *Air gaps: (a) Distributed-gap MPP; (b) Gapped ferrite.*

Two 920mH inductors



Different cores

moly permalloy powder

distributed air gap

sharp fall core saturates

Fig. 3. *Inductance roll off curve for two 920 mH inductors using MPP and gapped ferrite.*

L(i) curve shape

$I_L \downarrow$ $L \uparrow$

$I_L \uparrow$ $L \downarrow$

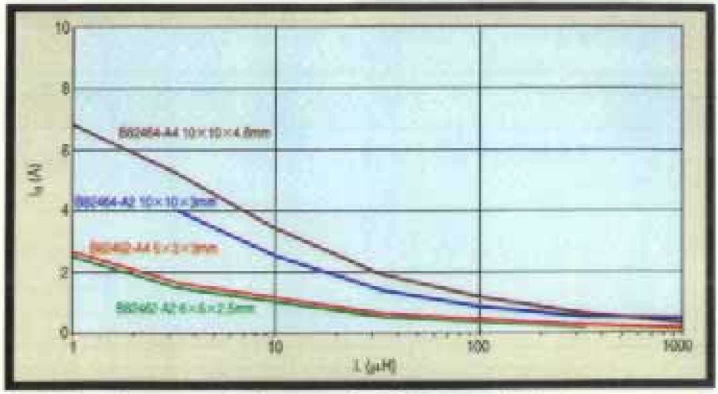


Fig. 7. *Rated current vs. inductance for SMD power inductors.*

The MAX of the Current Limit

The MAX of the current limit is not completely irrelevant. As we move to *higher and higher input voltages,* this may become worth watching out for. The reason is that whenever we start-up, or submit the converter to sudden transients, the current no longer stays at the steady value it has under normal operation (when delivering the required maximum load current). For example, if we suddenly short the output, the control, in an effort to regulate the output, expands the duty cycle to the highest permissible value. The current then ramps up to the current limit.

Ouch

But that means the inductor could be saturating!

For example, if we are using a 5A fixed current limit buck switcher IC for a 3A application, we have probably picked an inductor rated for around 3A. But when we short the output, the current momentarily hits the current limit (which may be around 5.5A nominal). And that clearly exceeds the rating of the inductor.

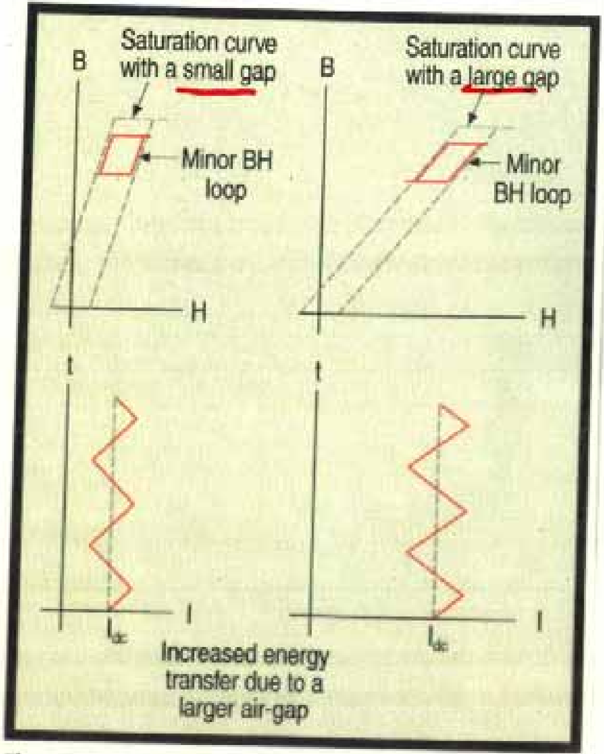
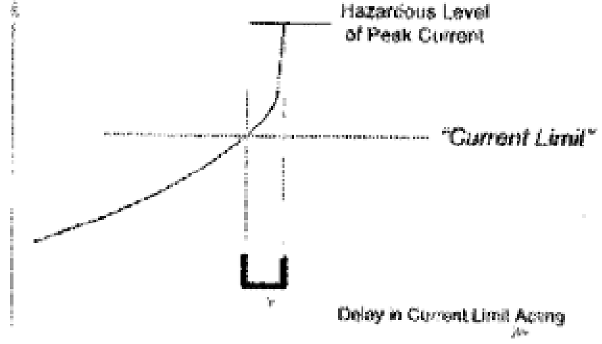


Fig. 2. *Impact of a gap in magnetic core material.*

But luckily, in most *low-voltage* applications, this doesn't cause any problem. Because, if the switch is rated for 5.5A, and the current limiting circuit in the IC acts ***fast enough*** to prevent the current from ever rising beyond that value, then even if the inductor has started saturating as it gets to 5.5A, there is no cause for concern --- after all, if the switch doesn't break, we don't have a problem! And since the current doesn't exceed 5.5A, the switch cannot break. So in this case, we could certainly pick a cost-effective "3A inductor" for our application, knowing fully well that it would saturate somewhat under various non-steady conditions.

Saturating Inductor Current Waveforms



The problem generally starts at input voltages ***higher than about 40V.*** Then the small delay between current limit being reached and the switch turning OFF, can prove fatal (see figure above). The ramp up of the fault current is *very steep* now as it is driven by

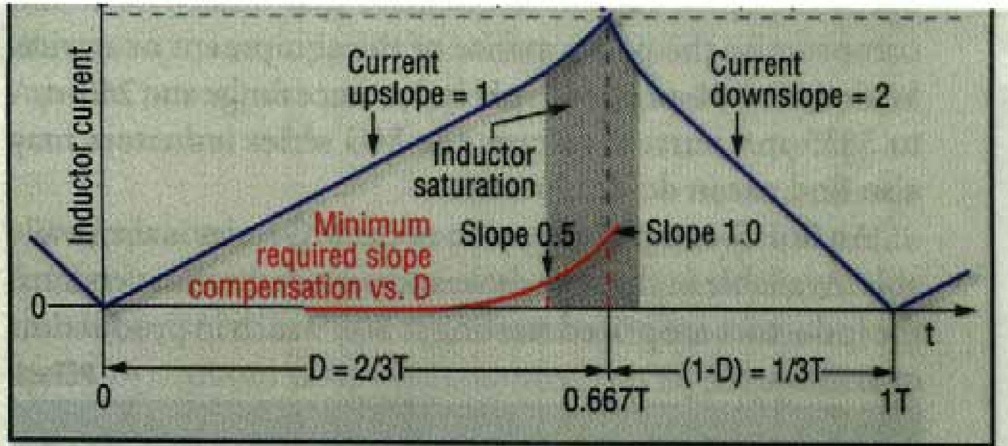


Fig. 2. *Slope compensation for inductors entering saturation is determined by the current upslope and downslope within the saturated region.*

DO3340 Series

Inductance 10 - 1000 μH

Isat up to 8.0 A, Irms up to 3.5 A

Kit C110 $60

DO3316HC Series - High Current

Inductance .33 - 4.7 μH

Isat up to 20 A, Irms up to 16 A

Kit C126 $45

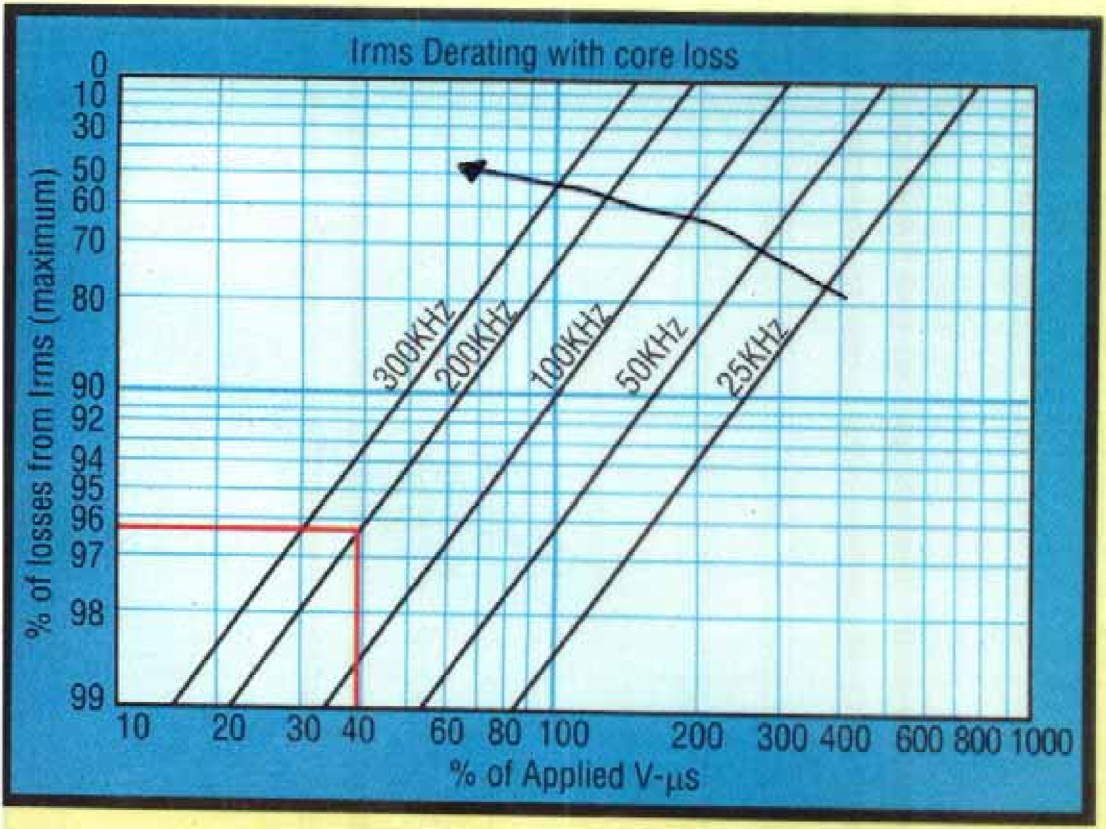


Fig. 4. *DR73 Irms De-rating Curve*

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20
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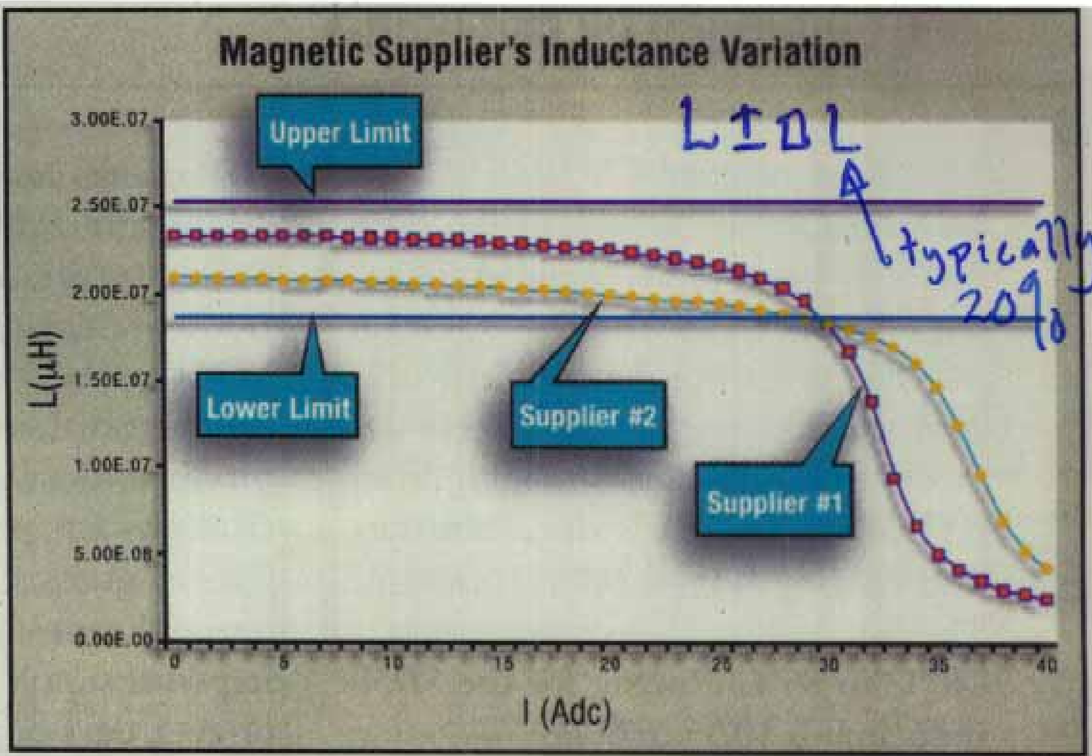


Fig. 5. *Inductor tolerances vary from one manufacturer to another, which can affect the design of the voltage regulator's error amplifier.*

Spreads and Tolerances when Selecting Inductance

We have a certain stated MIN and a MAX for the current limit (in the datasheet) This range (usually) includes process variations as well as variations over temperature. But the question is --- which of these limits should we consider for choosing the Inductance?

- **To guarantee output power, we need to consider only the MIN of the current limit** . In most low-voltage DC-DC converter applications, that is all that counts --- we can usually ignore the MAX (or TYP). So to guarantee output power, *we must ensure that the peak current in our application is certainly less than the MIN value of the current limit* .

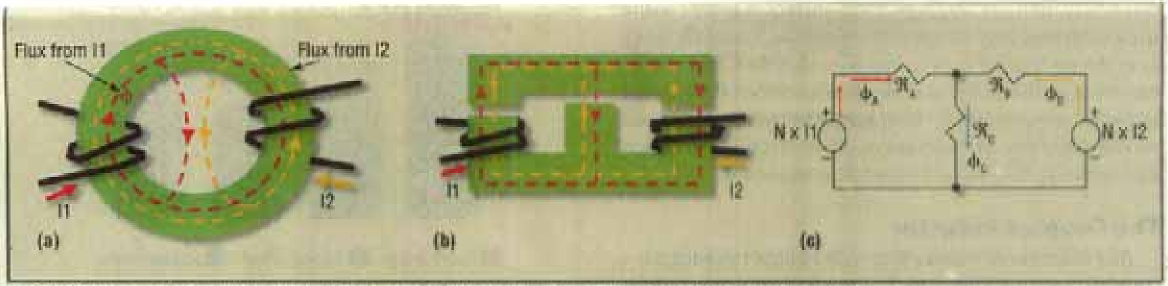
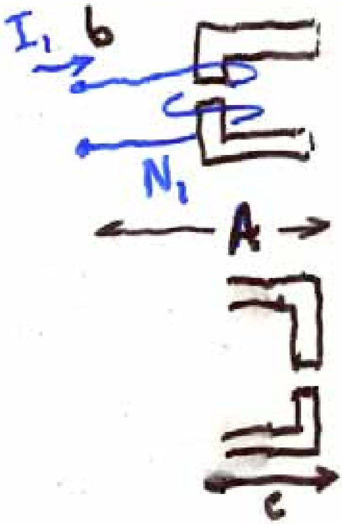
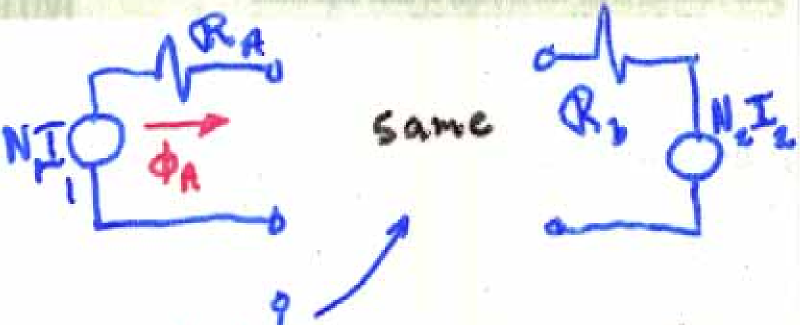
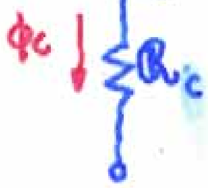


Fig. 2. *Both the toroid design (a) and the E-core design (b) can be reduced to the same reluctance model (c).*







$$\approx \frac{l_c}{\mu_c A_c} + \frac{l_{gap}}{\mu_0 A_c^2}$$

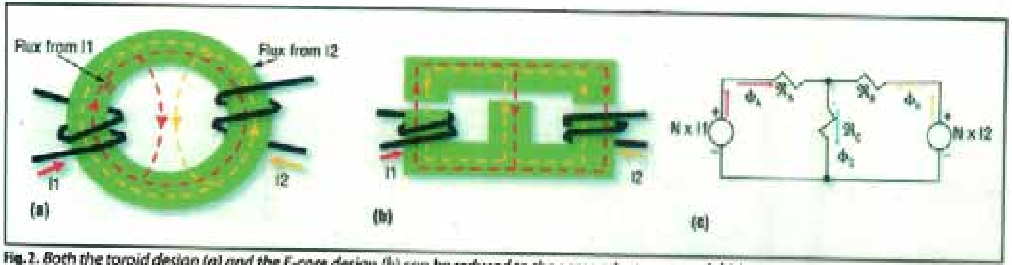


Fig. 2. *Both the toroid design (a) and the E-core design (b) can be reduced to the same reluctance model (c).*

$$MMF = NI = \sum_x \phi_x \mathcal{R}_x \, , \; \mathcal{R}_x = \frac{l_x}{\mu_x A_x}$$

Similiar to "K's" loop equations

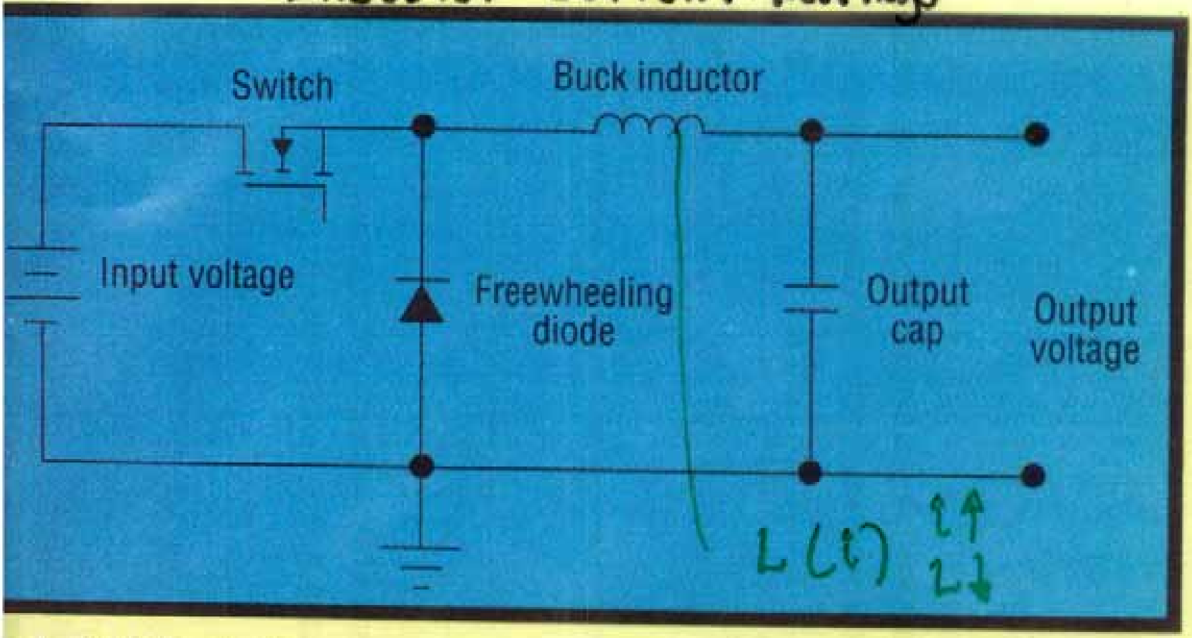
(a) $N_1 I_1 + N_2 I_2 = \phi_c \mathcal{R}_c$

$\uparrow \rightarrow 0$ for $\mu \rightarrow \infty$

$\mu = \infty$ $\quad I_2 = \left(\frac{N_1}{N_2}\right) I_1$ ideal trf

$\mu \neq \infty$

Inductor Current Ratings

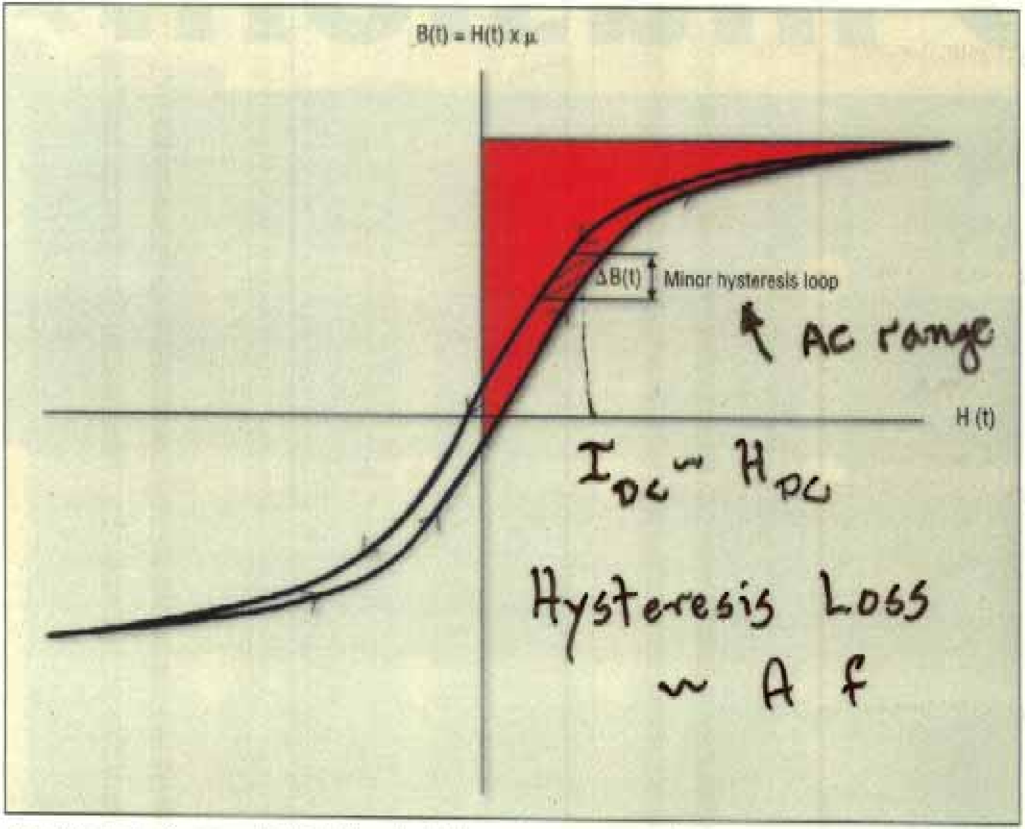


g. 2. Buck Converter.

$I_{rms} \longleftrightarrow i$ to cause $T_{core} \approx 40^{\circ}C$

$I_{peak} \rightarrow$ Limit "rolloff" of $L(i)$ to 20%

core saturation



↑ AC range

$I_{DC} \sim H_{DC}$

Hysteresis Loss $\sim A\,f$

Fig. 2. *A plot of magnetic field density B(t) versus magnetic field strength H(t) reveals the major and minor hysteresis loops associated with an inductor core.*

INDUCTOR POWER LOSS

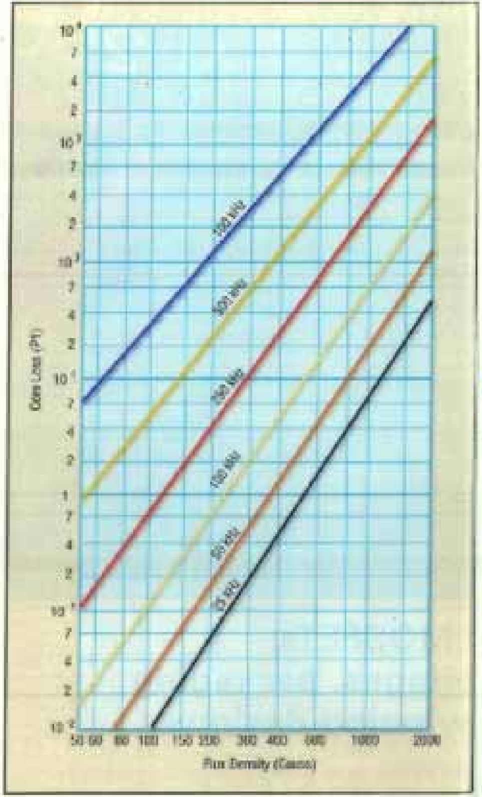
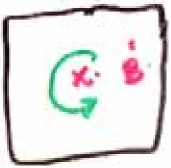


Fig. 1. *AC core loss for a particular ferrite material is plotted as a function of flux density at different frequencies. (Data courtesy of Spang and Co.)*

eddy current



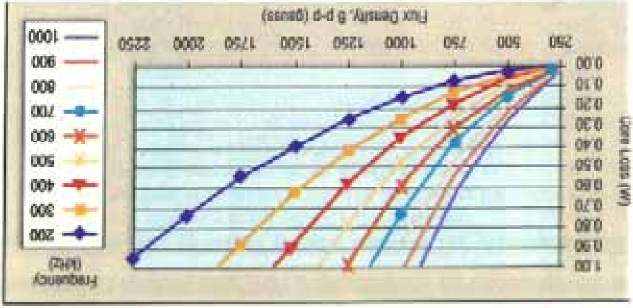
$$i_{ec} \cong V / R_{core}$$

$$V = w B$$

$$P^{ec}_{loss} = i^2 R_{core}$$

$$= \frac{V^2}{R_{core}}$$

$$P^{ec}_{loss} \sim w^2$$



INDUCTOR POWER LOSS

Recore of ceramic

A graph
is shown in
matters. It t
hysteresis lo
ΔB using th

$$\Delta B=K$$

where K
our case), a
example:

$$\Delta B(t)=$$

As an alt
ductor volt-
and the cor

$$\Delta B(t)$$

Going t
f_{sw}= 1 MH
equivalent
in the induc
RMS voltag

	Power loss	Saturation	Cost
Silicon steel	High	High	Low
Metglas	Low	Medium	High
JFE super cores	Low	Medium	Medium
Nickle alloy	Low	Medium	High
MPP	Low	Low	High
Sendust	Medium	Low	Medium
Iron powder	High	Medium	Low
Ferrite	Low	Low	Medium
Nanocrystalline	Low	Medium	High

Table 2. *Core materials.*

Rou

Rec
wir

Foi

Litz

Table

**Pow*

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com
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If the inductance is doubled, for example, the ripple cur

Limits of f_{sw} ↑ for smaller C, L

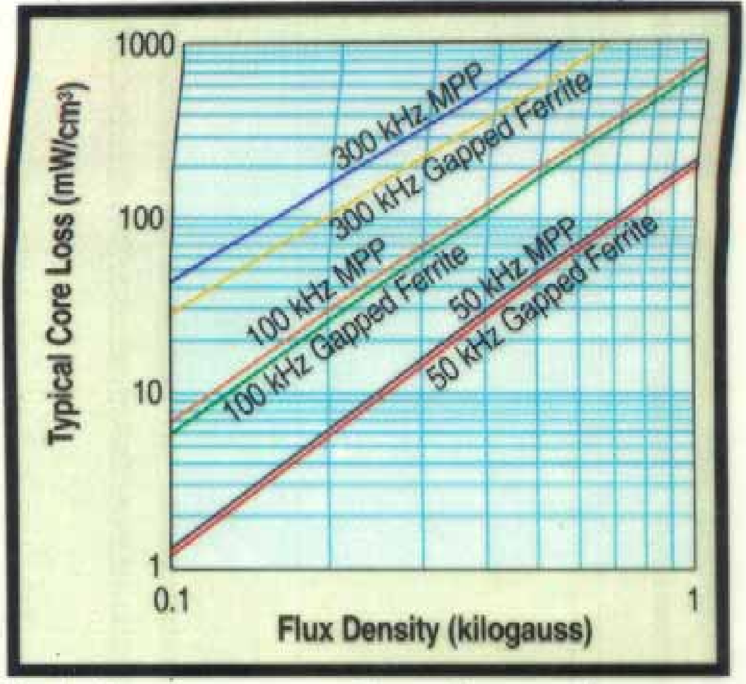
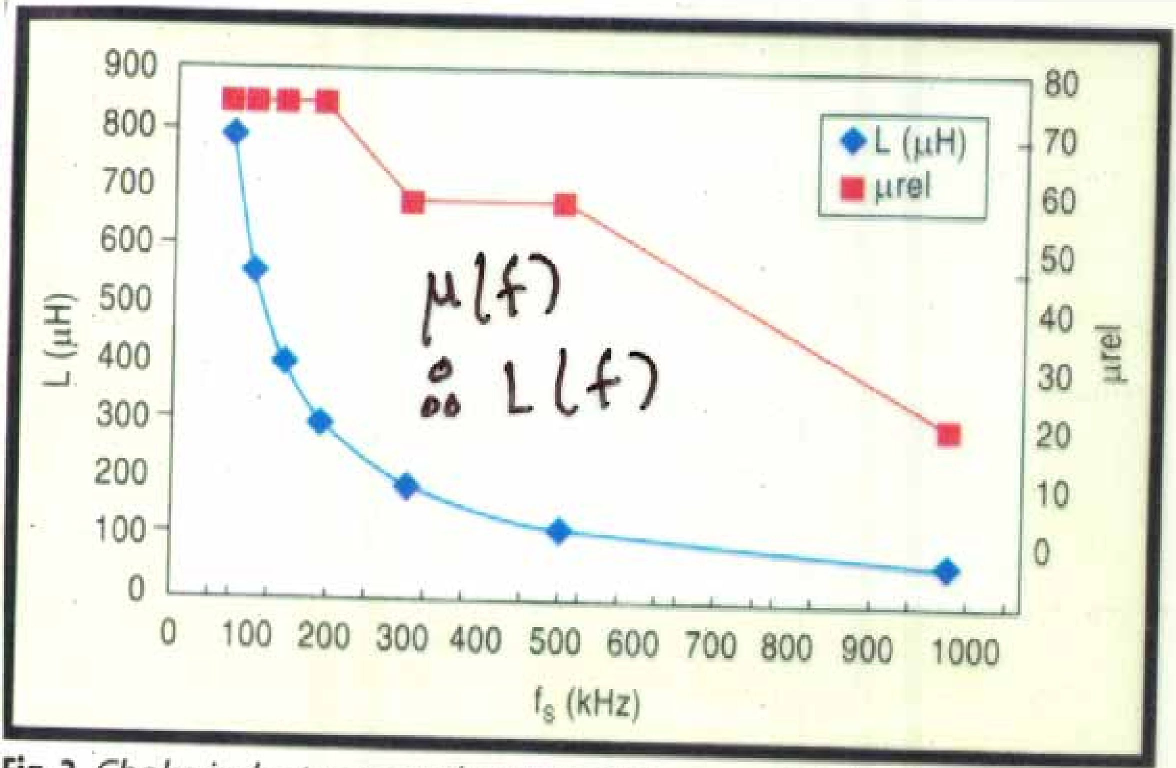


Fig. 6. *Core loss comparison of gapped core*



$\mu(f)$

$\therefore L(f)$

Fig. 2. *Choke inductance and permeability vs. switching frequency.*

wL vs f is ?

Coiltronics Part No	Rated Inductance at 1.8Apk (µH)	Dissipation (mW) (for 40°C rise)	De-rated Irms (A)	Core Loss (mW)	I^2R Loss (mW)	Total Loss (mW)
CTX10-1-52	7.5	277	1.98	89.4	108.3	198.7
CTX10-1A	8	371	2.7	35.7	103.5	139.2
DR73-100	9.5	282	2.03	11	142.7	153.7

Table 2. *Comparative Inductor Losses.*

Losses in <u>actual</u>
L Choices

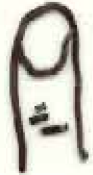
Q

D

Choosing Inductance based on Current Limit of the IC

Let us summarize the general procedure for selecting inductance for a switching power converter (later we will look at practical issues like "spreads" and "tolerances").

- We usually determine the inductance by requiring that the current ripple ratio is about 0.4 --- we know that r = 0.4 represents an optimum of sorts for the entire converter.



$\Delta I > .4$

Choosing a current ripple ratio much smaller than this will lead to an excessively larger-sized inductor, with no significant improvements in the various current stresses in other parts of the converter. On the other hand, a much larger current ripple ratio will give us a slightly smaller inductor, but with significant increase in the RMS currents (through the input/output capacitors in particular).

I_Q^{max}, I_D^{max}

Can ceramic take the I^2_{RMS} (ESR) loss

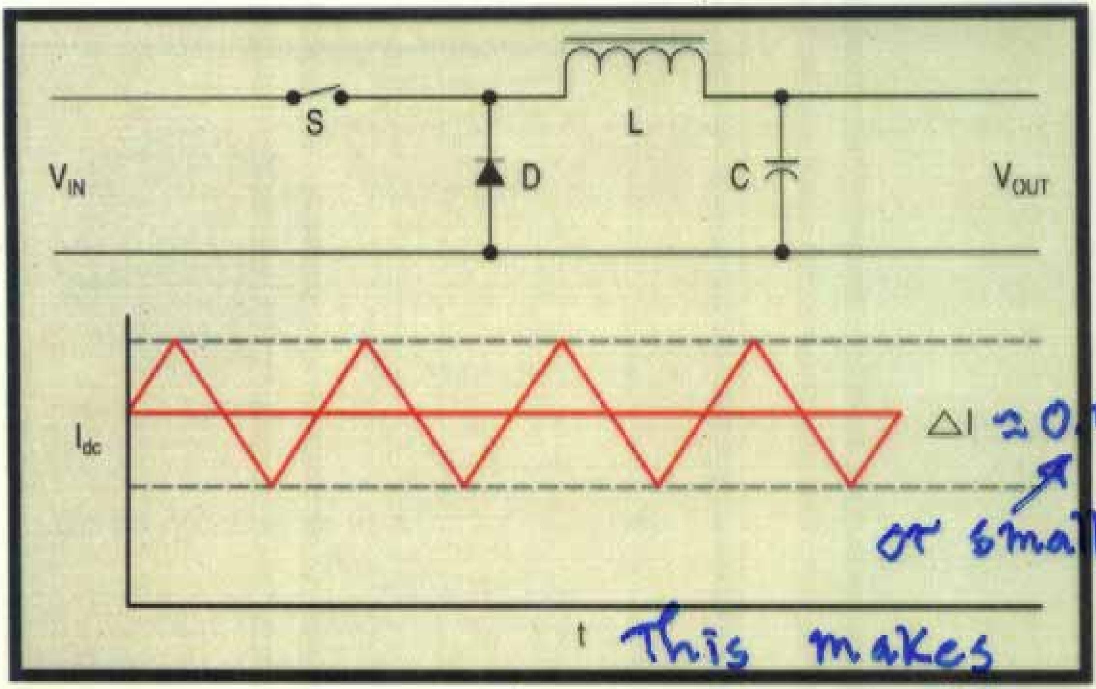


Fig. 1. *Buck converter.*

$\Delta I \approx 0.4 I_{dc}$

or smaller

This makes

$\Rightarrow I_{peak} = 1.2\ I_{dc}$

Minimum Rating of Inductor (theoretical) to ensure Power Delivery

So we know what the "rating" of a given inductor is. But how does it relate to the our application?

$r = \frac{\Delta i}{I_{dc}}$

In all converter designs, we need to start by setting the current ripple ratio "r" ---- defined as $\Delta I/I_{AVG}$. This indirectly determines the inductance (depending on the switching frequency).

Note that if "r" is set typically equal to say 0.4 (by suitable choice of inductance), the peak is then (by definition) 20% higher than the average inductor current. A current ripple ratio of 0.6 for example, means that the peak inductor current is 30% higher than the average inductor current. And so on...

The average inductor current in a given application is equal to

- Io (max load), for a buck.
- Io/(1-D) for a boost and buck-boost.

note I_L (DC)

We know that the peak current is by definition

$I_{PEAK} = I_{DC} + \frac{\Delta I}{2}$

$I_{min} = ?$

ΔI

$2\Delta i =$

So, in general, **if we choose the current rating of inductor to be higher than the calculated peak inductor current,** we should be fine --- at least from the power-delivery point of view.

I_L^{Peak} factors

For a buck that means,

$$\text{Irating} \geq \text{Io} \times \left(1 + \frac{\Delta I}{2 \times \text{Io}}\right) \text{buck}$$

$I_L \approx I_o$

and for a boost and buck-boost that means

$$\text{Irating} \geq \frac{\text{Io}}{1 - D} \times \left(1 + \frac{\Delta I}{2 \times \frac{\text{Io}}{1 - D}}\right) \text{boost and buck} - \text{boost}$$

$I_L \sim \frac{I_o}{1-D}$ (both boost buck-boost)

input circuit

V_{low} I_{low} (must go ↑ if V_o ↑

- But there is another consideration --- *if we are close to the current limit of the device, we need to ensure that the inductance is large enough not to cause the calculated peak current to be greater than the set current limit* --- otherwise in practice, "foldback" will occur and required maximum output power can no longer be guaranteed.

Avoid I_{sat}

For example , if we have a "5A buck switcher", being operated at 5A load, then *ideally* , we would want the current limit of the device to be at least 20% higher (current ripple ratio of 0.4). So we want $I_{CLIM} > 1.2 \times 5 = 6A$.

Unfortunately that much of margin (headroom) is rarely available in most commercial switcher ICs.

Rule

Choice of I_{sat} of chosen L

the *higher applied voltage*, and the peak current can therefore exceed the rating of the switch, thus eventually causing degradation and switch destruction.

The rule of thumb for selecting INDUCTOR SIZE is that

- **If the input voltage is less than 40V, pick an inductor with an I_{SAT} greater than I_{PEAK} of our application.**

- **If the input voltage exceeds 40V, for assured reliability, pick an inductor with an I_{SAT} greater than or equal to the MAX of the current limit** — for we know that as a worst case we may hit the MAX of the current limit.

Note that in fact, in off-line power supplies, all the magnetics are traditionally sized such that they are guranteed not to saturate under the MAX of the set current limit.

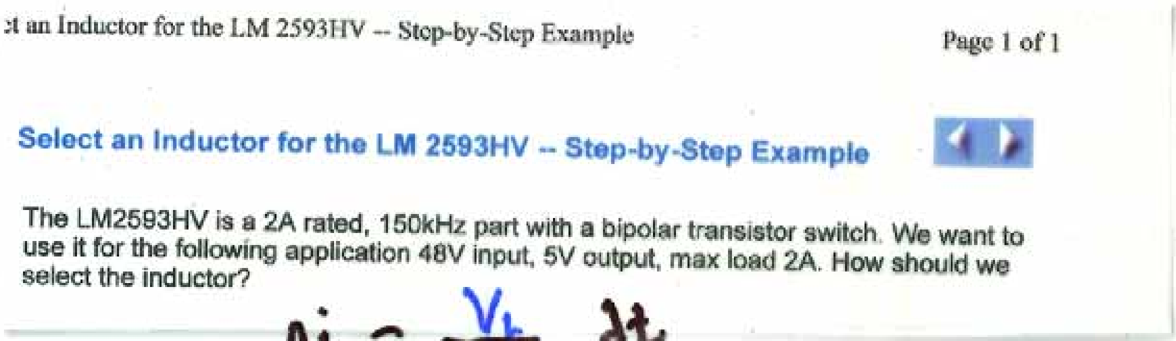
Note: Under high- or low -voltage conditions, the MIN of the current limit is still used to pick the inductance (power-delivery criterion).

(unsuccessfully) tried to exploit the larger size of the chosen core (by increasing the wire thickness), but the stumbling block may have been that the selected core geometry was somehow not conducive to doing so --- maybe it just did not have enough *window space* for accommodating the thicker windings.

Off

- In some inductors, we may even find I_{SAT} to be less than I_{DC}. What use is that? We can't operate beyond I_{SAT} in any case! So the only advantage, if any, in this case is that the temperature rise will be less than the maximum specified (when operated at I_{SAT}).

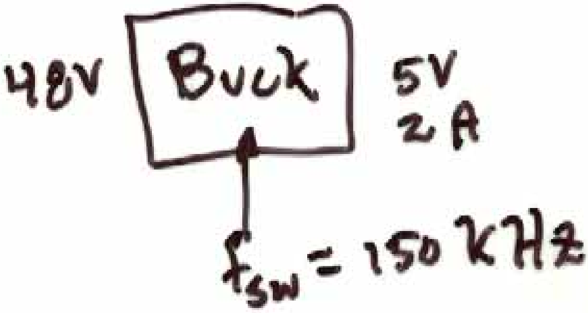
In general, for all practical purposes, the current rating of the inductor that we need to consider is the lowest rating of all the published current ratings. We can ignore all the rest.



Select an Inductor for the LM 2593HV -- Step-by-Step Example

The LM2593HV is a 2A rated, 150kHz part with a bipolar transistor switch. We want to use it for the following application 48V input, 5V output, max load 2A. How should we select the inductor?

$$\Delta i \simeq \frac{V_L}{L} dt$$



- Further, the value of inductance we thus calculate, is actually the MIN of the spread of inductance. Most inductors have a +,- 10% spread. Therefore the *nominal value of the inductance we choose must be about 10% greater than the calculated value of inductance* from the previous step!

- Note that, in addition, we would ideally like to leave about 20% headroom between the peak current of our application and the MIN of the current limit. This is intended to provide *good step-load and step-line response.*

But unfortunately, with fixed current limit switchers, this may be a luxury we can't afford --- because their MIN current limit is usually set only slightly higher than the rating of the device. To leave 20% headroom (assuming that is even possible), would demand a very large inductance (when operating close to the max load of the device). And we also know that a large inductance itself tends to slow down the transient response — just opposite of what we were hoping for here!

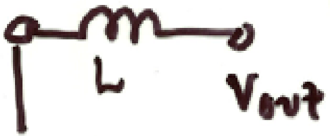
So in general, we usually end up ignoring this "nice to have" step-response headroom/margin (when operating close the load limits of the device), and concentrate on just assuring "power-delivery" ---- by simply requiring the MIN of the current limit to be higher than the peak current of our application, as indicated above.

$L \uparrow$ transient response suffers

$L \downarrow$ Could have Δi saturate core

Step	Instruction
1	Time Period (switching frequency fsw is 150kHz) $T = \frac{1}{fsw} \Rightarrow 6.67\mu s$
2	Duty Cycle (ignoring forward drops) $D = \frac{Vo}{Vin} = \frac{5}{48} \Rightarrow 10.4\%$
3	Off-time $toff = (1 - D) \times T = \frac{100 - 10.4}{100} \times 6.67\mu s = 5.98\mu s$
4	Voltuseconds $Et = Voff \times toff = Vo \times toff = 5 \times 5.98 = 29.9V\mu s$

D = 0.1 nominal



be aware Lmin for this condition of V-sec

5 $r \cong 2\frac{\Delta i}{I}$	Check limit of current ripple ratio 'r' at maximum load current. I_{CLIM}(MIN) is 2.3A. $I_{DC} \cong 2A,\ I_{max} = 2.3$ $\Delta i\ \%$ $r\ \lim it = 2\times\left[\frac{I_{CLIM}\ (MIN) - Io}{Io}\right] = 2\times\frac{2.3-2}{2} = \frac{0.6}{2} = 0.3$
6 Default	Select 'r' for application: if rlimit (calculated above) is less than ideal of 0.4, select r=rlimit, otherwise pick r=0.4. Therefore in this case we select r=0.3 (to guarantee output power)
7 $e = L\frac{di}{\partial t}$ $L = \frac{e\,\partial t}{\Delta i}$	Calculate MIN value of inductance using general equation LxI=Et/r (where I is the average inductor current --- equal to Io for a buck, and Io/(1-D) for the remaining) $L(MIN) = \frac{Et}{I\times r} = \frac{\overset{V\mu s}{29.9}}{2\times 0.3} = 49.83\mu H$ Δi bigger L better!
8	Pick Calculated Nominal Value of Inductance (assume final inductor will have +,- 10% tolerance --- so pick nominal value 10% higher than calculated value above). $L(NOM) = 1.1\times L(MIN) = 1.1\times 49.83 = 54.8\mu H$

$$e = L\left(\frac{di}{dt}\right)$$

$$L_{MIN} = \frac{e\,dt}{di}$$

Buck "L" for µP

SER2000 Series
Inductance: 0.3 - 2 µH
I sat up to 100 A
Designer's Kit C174 $60

100 A
<u>no</u>
Saturation

SER1590 Series
Inductance: 0.3 - 1 µH
I sat up to 50 A
Designer's Kit C166 $60

SER1360 Series
Inductance: 0.33 - 10 µH
I sat up to 40 A
Designer's Kit C165 $65

RoHS
COMPLIANT

available from vendors

9	Choose next highest standard inductance value. So in this case we choose L=56uH
10	Current Rating of Inductor --- if the input voltage was less than 40V, we would pick a current rating of (1+r/2)xIo=1.15x2=2.3A. However, in this case, since input is greater than 40V, we pick the inductor rating as per the MAX of the current limit i.e. 4A (from datasheet)
11	Final selected inductor is 56uH, 4A.

/www.national.com/AU/design/courses/252/ind03/07ind03s.htm

Coiltronics Part No.	Irms (A)	Isat (A)	DCR (mΩ)	Volt-μseconds rating (100 kHz)	Size/Volume (mm/mm³)	Core Material	Core Shape
CTX10-1-52	2.4	2.1	48.1	5.4	5.6 dia. x 4.7 / 272	Iron Powder	Toroid
CTX10-1A	2.84	2.5	46	10.25	8.89² x 4.19 / 331	Amorphous	Toroid
DR73-100	2.11	2.47	53.4	11.5	7.6² x 3.55 / 205	Ferrite	Shielded Drum

Table 1. *Inductor Options for Buck Converter Example.*

Some other V-sec
Conditions

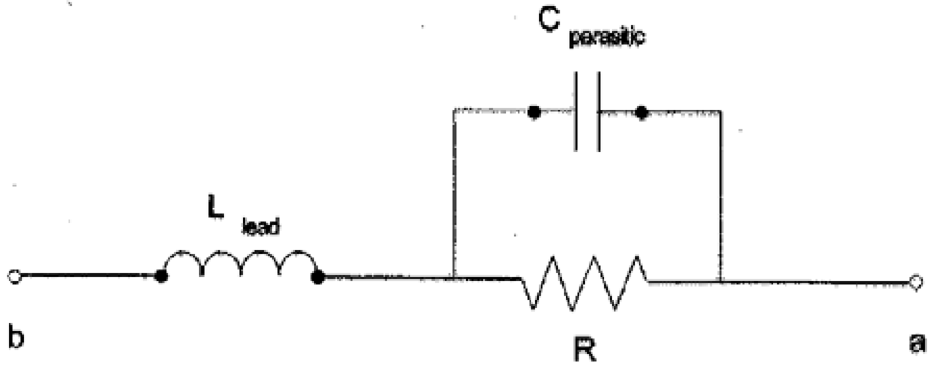


Figure 7. Simplified equivalent circuit for a resistor.

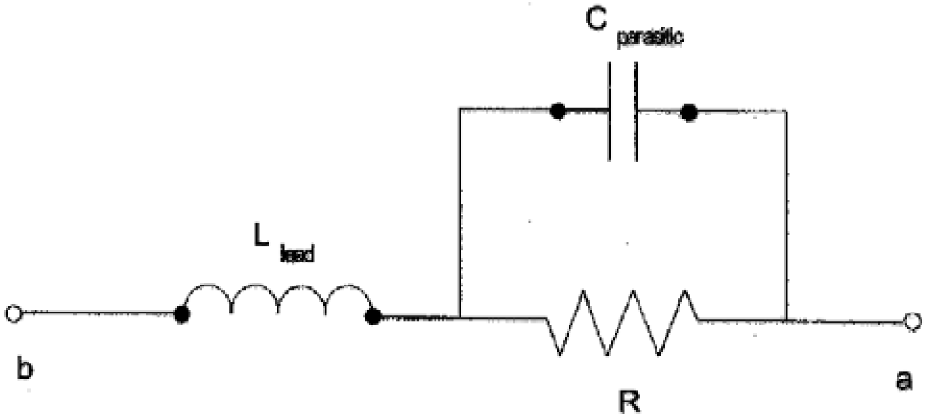


Figure 10. Equivalent circuit for resistor near self-resonance.

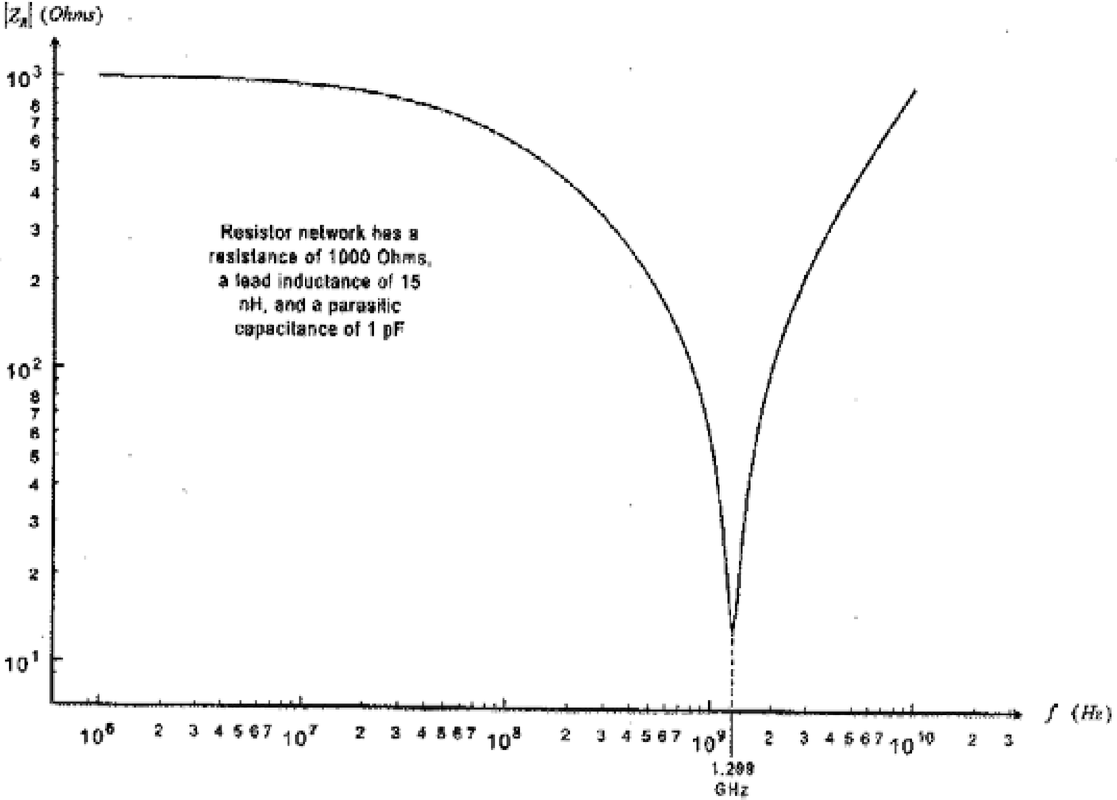
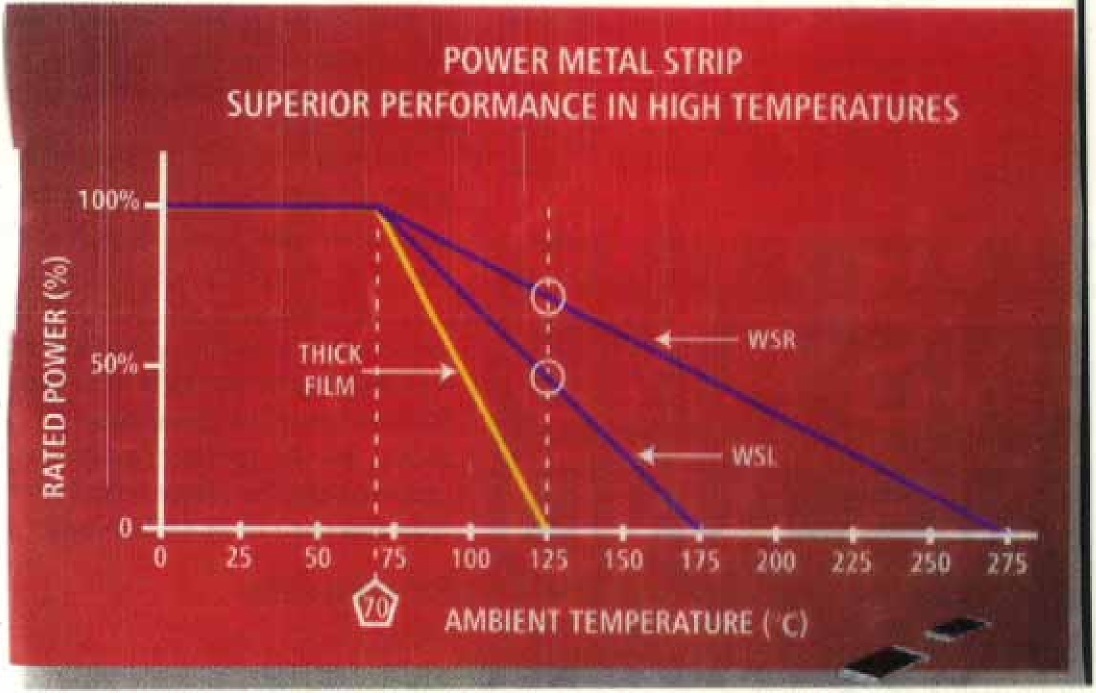


Figure 12. Frequency dependent behavior of equivalent circuit for 1000 ohm resistor.

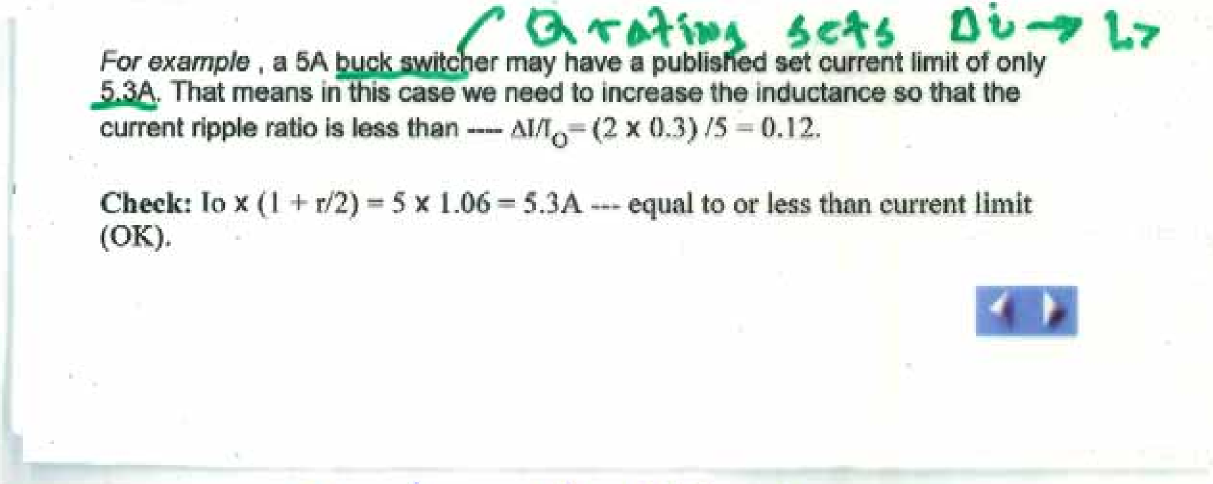
CURRENT SENSE TECHNOLOGY

for commercial applications

Need to "derate" power vs T



Q rating sets $\Delta i \rightarrow L \nearrow$



For example , a 5A buck switcher may have a published set current limit of only 5.3A. That means in this case we need to increase the inductance so that the current ripple ratio is less than ---- $\Delta I/I_O = (2 \times 0.3)/5 = 0.12$.

Check: Io x (1 + r/2) = 5 x 1.06 = 5.3A --- equal to or less than current limit (OK).

I_{DC} (max for Q) →

I^{peak} (max for Q) set by $\frac{\Delta i}{I_{DC}} \sim \frac{1}{L}$

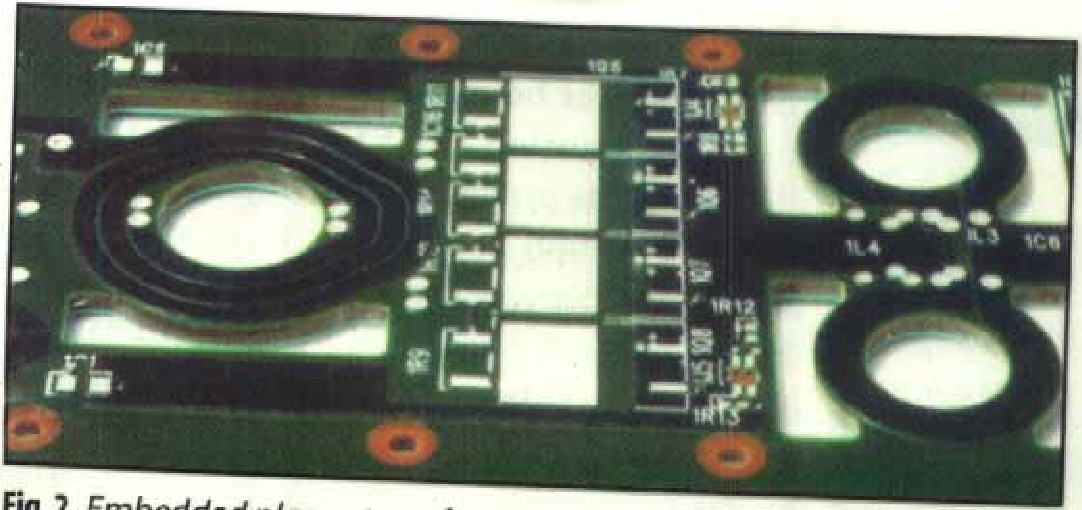
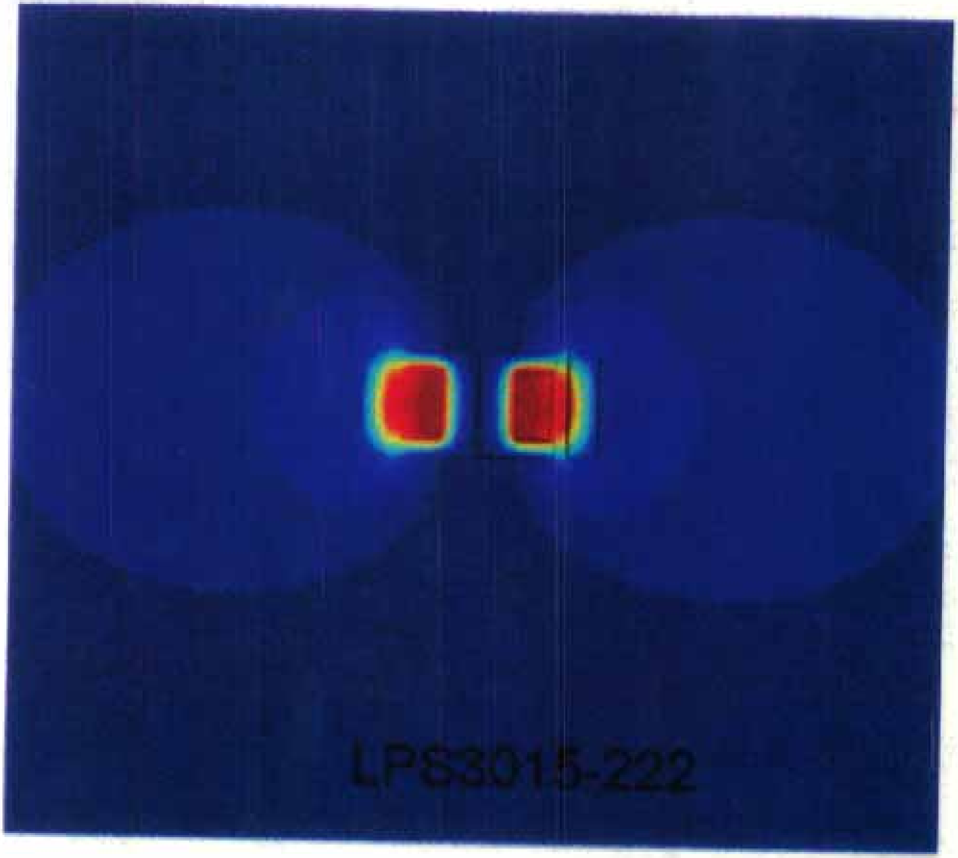
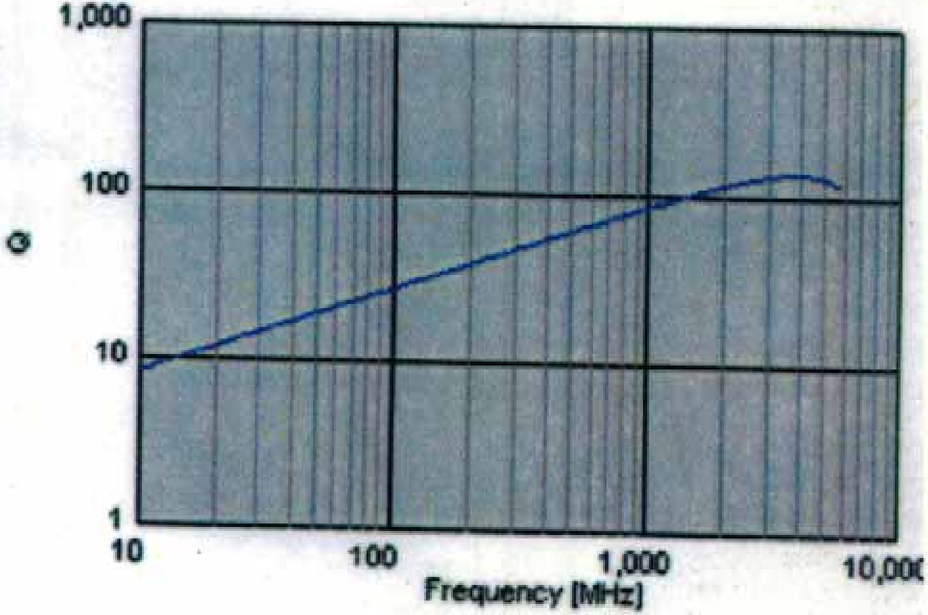


Fig. 2. *Embedded planar transformers.*

- Shielded Inductor Flux:





(b)

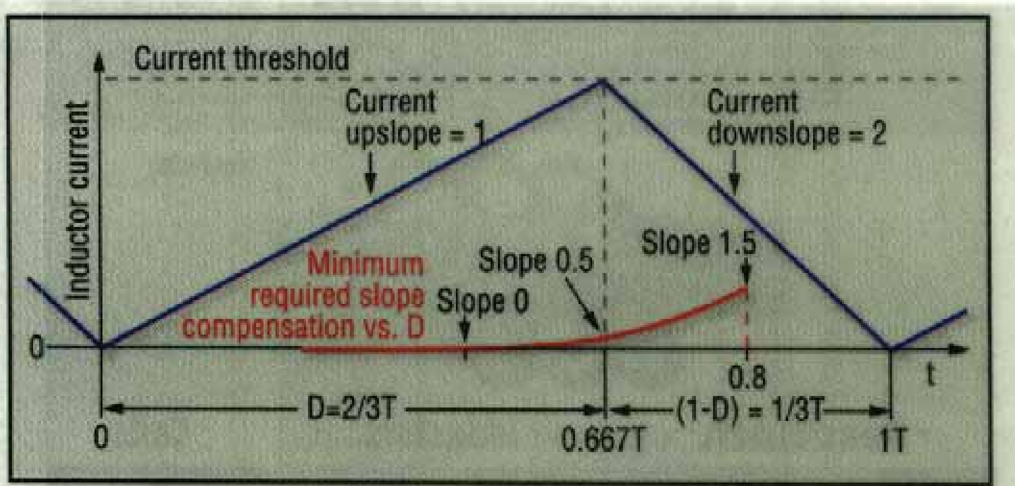


Fig. 1. *The miminum required slope compensation for peak-current control of PWM switching power supplies increases with duty cycle.*

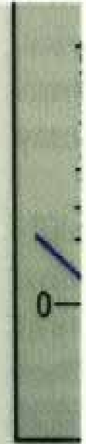


Fig. 2.
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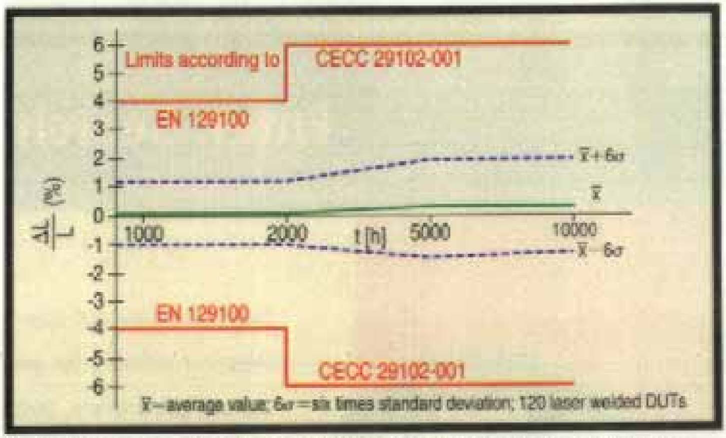
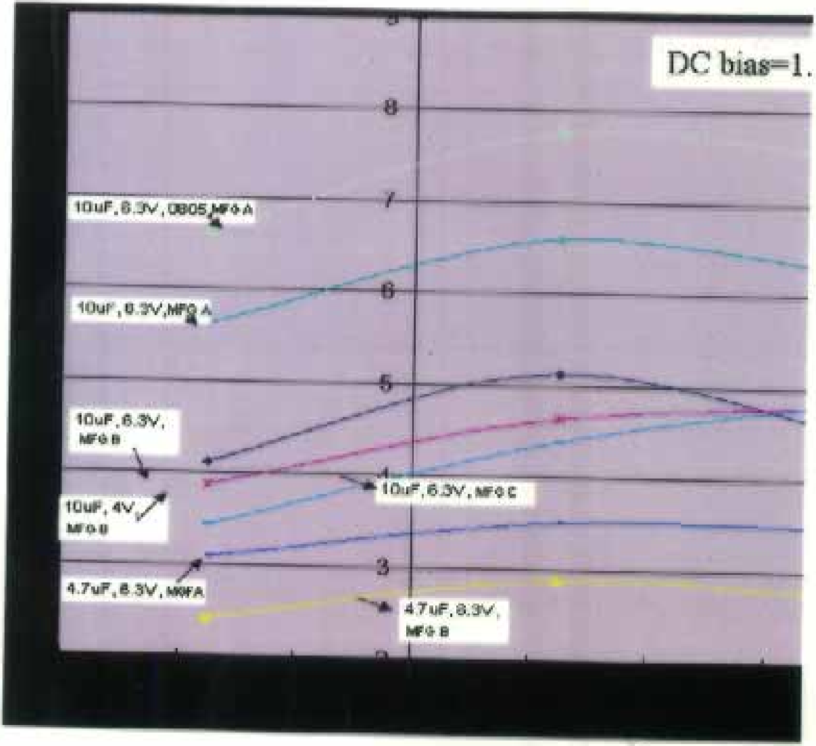


Fig. 4. *Inductance drift for rated current and ambient temperature of 125°C.*


DC bias=1.
8
7
6
5
4
3
10uF, 6.3V, 0805, MFG A
10uF, 6.3V, MFG A
10uF, 6.3V, MFG B
10uF, 4V, MFG B
10uF, 6.3V, MFG C
4.7uF, 6.3V, MFGA
4.7uF, 6.3V, MFG B

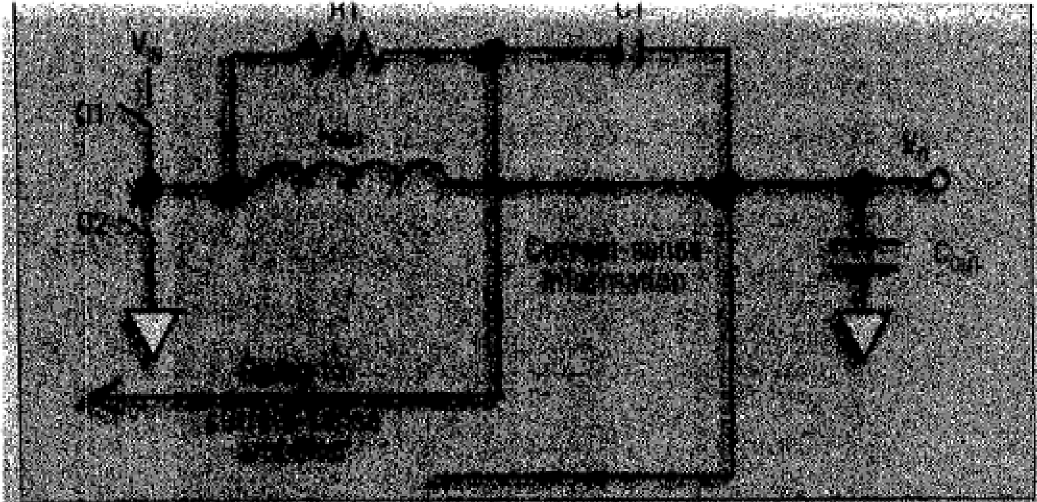


Fig. 2. *Replacing R_{SENSE} with an inductor for current sensing reduces power losses and increases the efficiency of the regulator. Voltage-sensing information is across C1.*

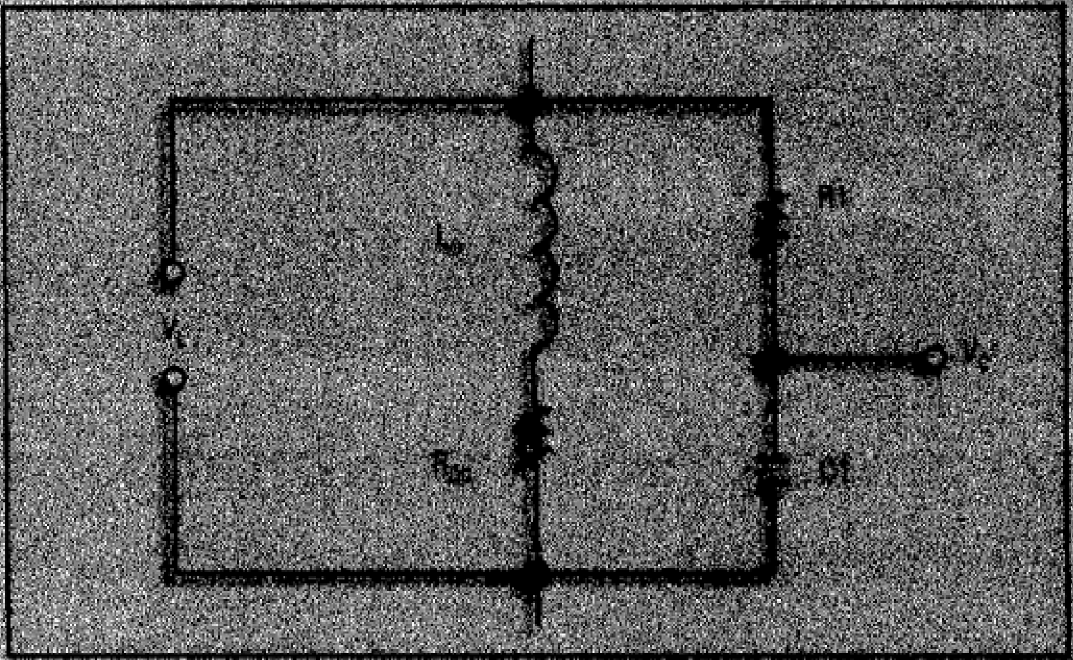


Fig. 3. *In inductor current-sensing, the voltage across the inductor and its internal dc resistance is the same as across the R1C1 filter network. This allows an equation to be written for the current-sense output across C1 in terms of the dc resistance of the inductor.*

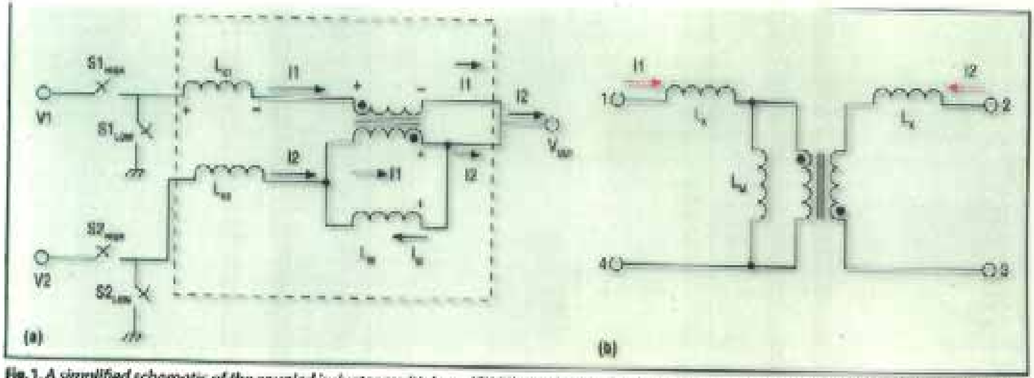
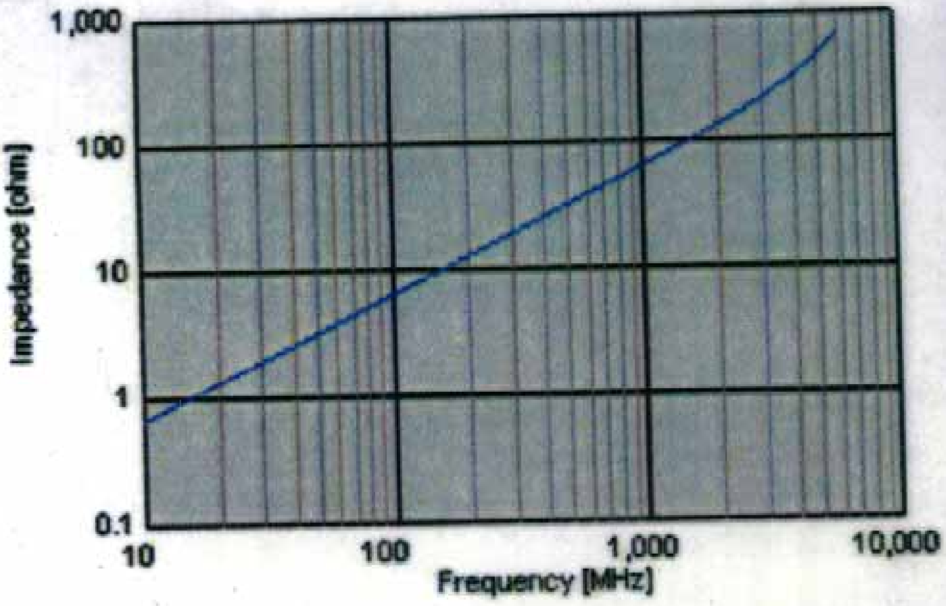
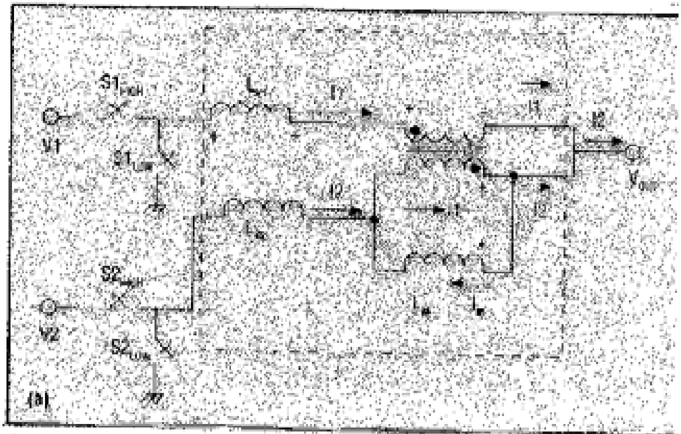


Fig. 1. *A simplified schematic of the coupled inductor multiphase (CIMP) topology is shown (a). The coupled inductor schematic (b) is identical to that of any two-winding transformer. The equivalent resistive losses (due to core loss and copper loss) as well as interwinding capacitances are ignored for simplicity. The magnetizing inductance (L_M) can be placed on either the primary or secondary winding with no effect to the*



(a)

Figure 3 Fre



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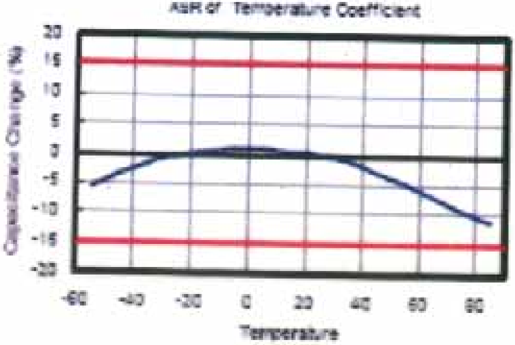
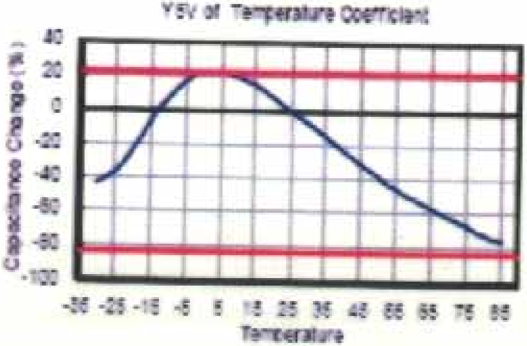
Any Inductor has Stored Energy --- even parasitic inductances

Once we understand an inductor, we know that it represents **stored energy** --- equal to ½ x Lx I^2 . And that is why we provide a "freewheeling path" for the main inductor. However, *even parasitic inductances have stored energy*. What about their stored energy? Don't we need to provide a freewheeling path for them too --- or suffer the consequences??!

f_{sw} ↑ "L" ↓ ⇒ parasitic effects L_{par} ↑





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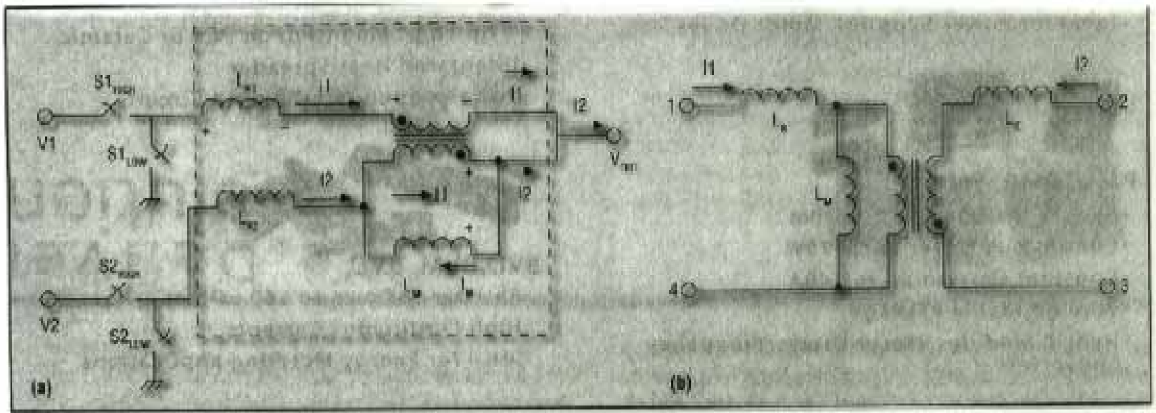

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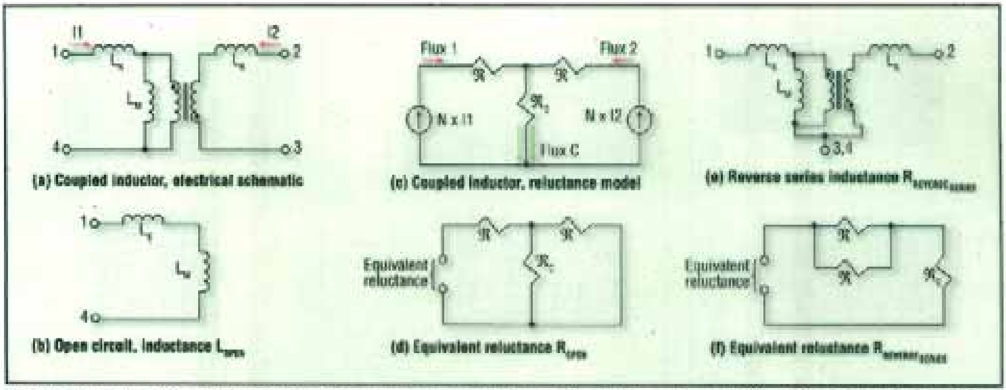


(a) Coupled inductor, electrical schematic

(b) Open circuit, inductance L_{OPEN}

(c) Coupled inductor, reluctance model

(d) Equivalent reluctance R_{OPEN}

(e) Reverse series inductance $R_{REVERSE_{SERIES}}$

(f) Equivalent reluctance $R_{REVERSE_{SERIES}}$

Fig. 3. *In order to write the equations for the circuit elements L_M and L_L in terms of the reluctance elements ℜ and $ℜ_C$, it is helpful to analyze*

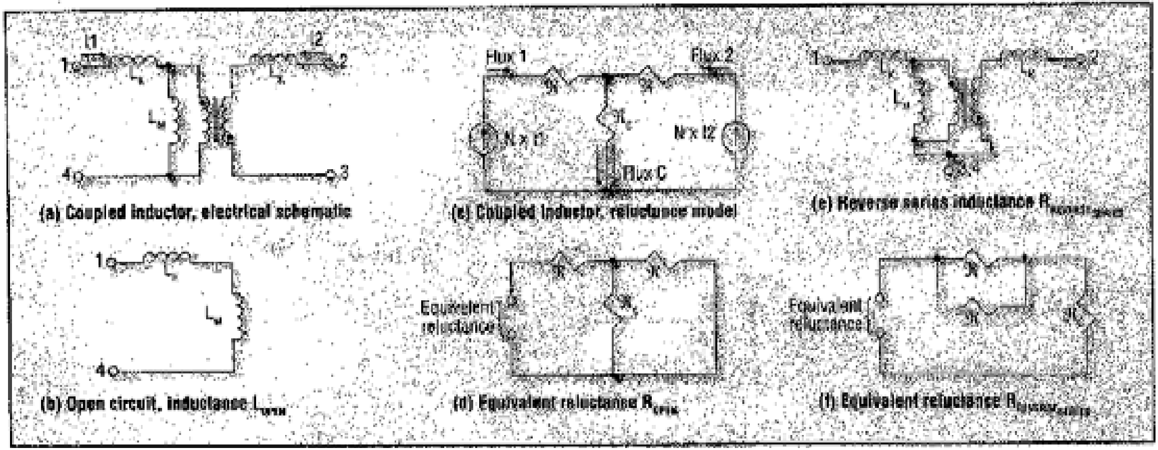
Independent Inductors	IMVP6 Notebook Specification	Coupled Inductors
	Topology: Two-Phase Buck V_{IN} = 12 V to 19 V Frequency = 300 kHz V_{OUT} = 1.2 V I_{OUT} = 44 A dc	
PG0255.401NL	Part Number	Pulse
2	Quantity/Board	1
360	Inductance (nH) - L_x	676
N/A	Inductance (nH) - L_M	183
1.0	DCR (mΩ)	1.0
11.5 x 10.3	Footprint (mm max)	18.4 x 8.8
4.0	Height (mm max)	4.0
10	Phase Ripple (A pk-pk)	10
30	di/dt max (A/μs)	60 (2x increase)
237	Total Board Area (mm^2)	162 (32% reduction)
2.0 (2 x 1.0)	Total Relative Cost	1.3 (1 x 1.3) (35% decrease)
1.7 (2 x 0.85)	Power Loss in Core (mW)	0.063

Table. *Two-phase coupled inductor design versus uncoupled two-inductor solution.*

$\phi_3 = \phi_1 + \phi_2$, (Eq.12)

where I1 and I2 are the currents in phase

A potential application for the CIMP topology



(a) Coupled inductor, electrical schematic

(c) Coupled inductor, reluctance model

(e) Reverse series inductance $R_{reverse series}$

(b) Open circuit, inductance L_{open}

(d) Equivalent reluctance R_{open}

(f) Equivalent reluctance $R_{reverse series}$

Fig. 3. *In order to write the equations for the circuit elements L_M and L_X in terms of the reluctance elements $\Re$ and $\Re_C$, it is helpful to analyze several possible inductance measurements and how these measurements affect the reluctance model.*