LECTURE 31
Inductor Types and Associated Magnetic Cores

A. Magnetic Core Choices to Wind Cu Wire Around
   1. Available Magnetic Cores
      a. Commercial Core Geometry's
      b. Available Core Materials(B_{SAT},μ vs. f, etc)

   2. K_g: Inductor Core Design Parameter

   3. **Magnetic Cores**
      a. Four Conflicting Needs in Magnetic Cores
      b. Specific Features of Selected Cores
         1. EE
         2. EC
            ETD
         3. PQ
            RN
         4. Pot Cores

B. Inductor Types: Depends on Required Current I_L(waveform)
   1. Filter/Coupled Inductors (Big I_{DC} and small I_{DC})
   2. AC Inductors (I_{DC} = 0 and I_{AC} only)

For nice web sites on commercial inductors see: www.coilcraft.com or www.premmag.com

LECTURE 31
Inductor Types and Associated Cores

A. Magnetic Core Choices

Inductors are made, by winding copper wire around magnetic cores. The cores usually contain an air gap purposefully cut into them to improve energy storage. Since the role of an inductor is to store energy, we will usually have one or more air gaps in the magnetic flux path of the core employed for an inductor. These air gaps will be precision machined as specified by the user. Air gaps help avoid exceeding B_{SAT} and also reduce B_{MAX} to further reduce core losses. There are a variety of core materials and many core geometeries for making inductors that operate at various frequencies and different current levels. The choice between the design possibilities revolves primarily around the specific circuit application needs that specify the value of inductance needed and the inductor current waveform that must pass through the inductor. In second order the core size and core shape chosen depends on **the area of the core winding window** as compared to the area of copper turns required, the thermal heating problems envisioned and electromagnetic interference issues that may arise from rf currents in the wire coils.

In summary, inductors have only one Cu winding around the magnetic core and the purpose of the core is to store energy. Transformers, while at first blush similar to inductors, have at least two Cu windings around the magnetic core and the purpose of transformers is solely to transfer energy between several windings with minimal energy storage. Keep this in mind.

1. Core Shapes Available

Appendix 2 of the Erickson text has five major magnetic core types listed for instructional purposes. Generally, available cores differ primarily in the core geometry, which has two major parts: the geometry employed for the magnetic flux paths and the open-air region of the core where Cu wire is wound around the core. Air gap location and size of the air gap, l_g is a third and separate issue. To achieve the required inductance we need sufficient room in the core wire winding window both for the copper turns and the required magnetic reluctance of the flux path in the core itself. $L=N^2/\text{Core Reluctance}$.

For cost reasons one should choose the smallest volume core that does the required job. This means a slight preference for POT/EC/EE/PQ/ETD core types. Core saturation on the other hand, at B_{SAT}, requires the largest core area for flux paths so that the catastrophe of core saturation does not ever occur. Thermal heating limitations of the core by power lost in both the wire windings and in the core itself will add additional constraints to the required core geometry needed to achieve heat transfer to the ambient. **In short, we need to make design tradeoffs in choosing inductor cores and this is the subject of lectures 31 and 32. In this lecture we do the design trade –off for filter inductors only.**

2. Inductor Core Design Parameter: K_g(inductor core)

A filter inductor design parameter, **K_g(inductor core),** is artificially created and used for comparing cores so that we may more clearly make the design tradeoffs required. One such core parameter is used only for inductor cores and is termed K_g(inductor). Another related but distinct core parameter, K_{gfc}(transformer), is used only for cores employed in transformers. Common to both inductor and transformer "K core design parameters" are terms like: MLT for the wire wound around that specific core geometry, the core area ,A_c , employed for magnetic flux paths,and the available window area ,W_A , for copper wire windings as shown below. Note that core parameters for transformers have additional terms compared to those for inductors. We will only focus on K_g(inductor) in this lecture. On page 18 **we will derive the relation for K_g(inductor)** by considering the four quantitative constraints on the filter inductor equations listed on page 18-19. For now we will just anticipate the **K_g**

results and compare to the transformer $\mathbf{K_{gfe}}$ factor to be introduced later.

Inductor Core | **Transformer Core**

$$K_g(\text{inductor}) = \frac{A_c^2 W_A}{MLT} \qquad K_{gfc}(\text{trf}) = \frac{W_A \ A_c^{2(1-1/b)} \ U(b)}{MLT \ \ell_e^{2/b}}$$

K_g is **solely the Geometric Constant** Unique to each core

K_{gfc} is the Core Property @ f we are operating and β~2.6-2.8

We isolate the **three major geometric factors for magnetic cores**:

1. $\mathbf{W_A}$ window area
2. $\mathbf{A_c}$ core area
3. **MLT** mean length/turn for the core

The top 3 geometric factors are present for both inductors and transformers. The additional terms for the transformer cores are

1. l_e : core magnetic flux path length
2. U(β) ≈ 0.3 for usual cores but will vary with each core type
3. β power law for magnetic loss in the core material chosen which for ferrites is in the narrow range 2.6-2.8

Including both hystersis and eddy current losses we find:

$$\frac{P}{V} \left(\frac{W}{cm^3} \text{ lost in core}\right) \sim K_{fe} \ B_{max}^{b} \quad \text{typically } 300 \text{ mW/cm}^3$$

$$P_{(\text{total in core})} = \frac{P}{V} * V_c (\text{core volume})$$

We must carefully distinguish between core loss per unit volume and total core loss. Each parameter will have its use in core choice for applications.

3. Choosing Proper Magnetic Cores

a. Four Conflicting Issues Arise: Heat Dissipation, Magnetic Saturation ,Wire Winding Needs and Air Gaps

1. Thermal Heat Dissipation Issues for Magnetic Cores

A core must be able to dissipate both wire winding and core losses to the ambient via both radiative and convective heat loss and still maintain a surface temperature below 100 degrees Celsius. We will show below that the required thermal resistance of cores to this dissipative heat flow, which we must design for, varies over the range: $0.10 \le R_Q(\text{core}) \le 100 \frac{°C}{W}$

In general for larger size cores the thermal resistance reduces allowing for higher heat flow with less temperature rise at the core surface. See the table for pot cores which includes thermal resistance, as shown on page 15. The allowable heat flow, Q, from the core, which is maintained at surface temperature, T_s, to the ambient at temperature, T_a, goes as Q x R(core) = T_s-T_a. Since we cannot allow $T_s \ge 100°C$, this sets the required R(core) for a given heat flow Q from both winding and core loss. In short, we need to calculate both winding and core losses expected for the circuit conditions and then design the core size and shape to meet the cooling needs as well as circuit needs. . Core losses act to increase T_s(core) which could lead to lower B_{SAT} as well as to wire insulation meltdown or worse.

2. Magnetics Limitation Issues

Generally Ampere's law relates the electrical and magnetic variables in a core. NI = Hl, that is for a given applied mmf the size of the magnetic flux path, l, (core size) may be increased to lower $(H)_{max}$. Lower H_{max} is important both due to core saturation and core losses as well as to the proximity effect in wire losses, which are placed in the core winding window where leakage magnetic fields flow. We want $B_{max} << B_{sat}$ to avoid core saturation and inadvertent circuit shorts causing possible switch and circuit catastrophes, but also to avoid large core and wire winding losses.

$$P\binom{\text{core}}{\text{loss}} = K_{fc}\ B_{max}^{b}$$

Choose magnetic materials accordingly to guarantee $B_{core} \le B_{saturation}$

2Testla	1T	½ Testla
Fe:Si	Powered Iron	Ferrite

Although cores may be the same for inductors and transformers, the

B(saturation) mechanism arises very differently in inductors and transformers as shown below due to the very different electrical circuits in the two cases. In particular inductor cores will saturate with large currents in the windings, whereas transformer cores tolerate very large wire currents but saturate because of DC offset voltages across the windings.

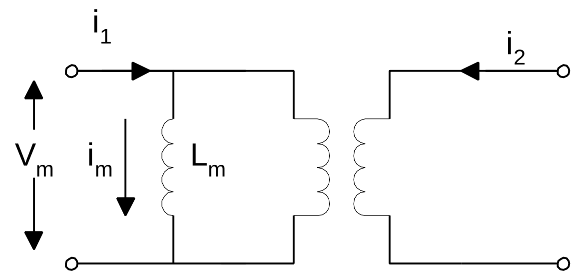
<u>Inductor:</u> B_L(Saturation)
Only one wound wire

$H_L \cdot l_c = N i_L$

$$H_L = \frac{N i_L}{\ell_c} \approx \frac{B_L}{m}$$

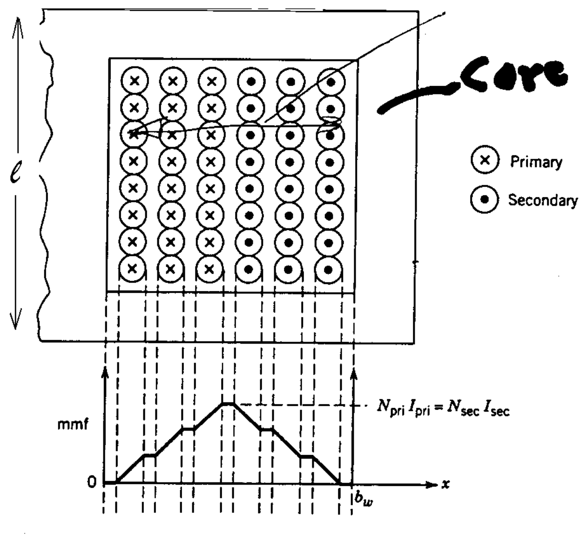
The maximum is set by:
$B_L \sim i_L$(max current)
$B_L \sim 1/l_c$ Flux path length of the core effects the maximum B

<u>Transformer:</u> B_m(saturation)
Two isolated wire windings with opposing mmf's



$i_m \neq f(i_1, i_2)$. **Current levels in the wire windings have no effects on B_{max} in the core**. I_m is set by volt-sec balance $i_m = \int V_m dt / L_m \sim B_m$ Thus we see, Volt-second balance sets the maximum B.

In transformers,leakage inductance's, L_l, store energy outside the core-in the air window where Cu wire is wound. L_l will <u>cause</u> unexpected V spikes as they discharge due to inductive kick. One can wind the coils in a special inter-digitated way to reduce L_l



In terms of energy in the primary:

$$\frac{1}{2}\ L_\ell(I_p)\ \equiv\ \frac{1}{2}\int m_o H_{w(airwindow)}^{\ 2}\, dV$$

To reduce the L(primary) we need to reduce H(air window). This is achieved by:

1. Core height is incresed so n turns fit in one layer with increased size l

   $Ni = Hl$ or $H = Ni/l$; $l \uparrow\ H \downarrow$

2. Interleave windings so mmf vs x has lower peaks.

3. Copper Wire Winding Issues

We choose the number of wire windings to achieve the desired inductance value. $L = N^2/R(\text{core})$. Wire windings must be wound around the core geometry chosen. The wire diameter chosen to handle the required current sets the number of wire turns that will fit in the area of the core set aside for the wire windings. Wire diameters are chosen based on the required DC and AC current levels that must pass through the wire. For complex current waveforms we can always calculate an RMS current.

$\frac{I_{rms}}{A_{cu\ wire}} = J_{rms}$. To keep wire winding loss equal for primary and secondary coils of a transformer, so one winding doesn't heat up prior to the other, we always seek to keep J_p(primary) = J_s(secondary). This means taking I_p and choosing A_p to get the desired J. We then match I_s with a wire with A_s to get the same J. For off the shelf AWG# wire this means picking wire sizes as close as possible to the equal J rule.

The number of turns of wire, with a given wire diameter, allowed in a core winding window depends on the core winding area. Wire with external wire insulation will take even more room and less turns will be allowed. K_{cu} takes into account the effective area of the wire devoted to copper. Thus we can say for an inductor the number of wire turns is set by our choices of core and wire size:

$$N \equiv K_{cu} \frac{A(\text{winding window})}{A_{\text{cu wire}}(\text{AWG\#})} =$$

The known inductor current sets the mmf which in turn sets the maximum flux in the core via the simple relation mmf=flux times magnetic reluctance of the core. From core cross-sectional area we find the corresponding B_{Max} which we compare to B_{Sat} of the chosen core. $f(\max) = B_{max} A_{core}$. Next we try to formalize this set of compromises for an inductor.

Lets make an aside for a homework problem. To emphasize the role of energy storage for an inductor we recall that the energy stored goes as ½ L i^2 and this is related to the core and winding parameters. To account for complex current waveforms we arbitrarily use the parameter shown below as follows:

$$LI(\max)\, I_{rms} = K_{cu}\, A(\text{winding window})\, A(\text{core})\, J_{rms}\, B(\max)$$
$$\downarrow \quad \downarrow \qquad \downarrow \qquad \downarrow \qquad \downarrow \qquad \downarrow \quad \downarrow$$

Energy stored in PWM circuit design inputs, L, I(max), I_{rms} — **Geometry of core and windings** — Core and Wire and Material limits

The material limits of copper wire maximum current densities and core saturation fluxes are well known constants. Circuit issues define the left hand side electrical parameters. Hence the core geometry's for both winding windows and flux path areas are determined. In short, we can determine the required core.

Let us assume a characteristic core size is set by one dimension, "a". This will allow for some crude rules of thumb for you to develop for homework.

To a first approximation for few layers of wire windings:

A(winding) * A(core) ≈ a^4; "a" characteristic size of core.

From J_{rms} and core size "a" we have to first order a relationship that you must prove for HOMEWORK # 1

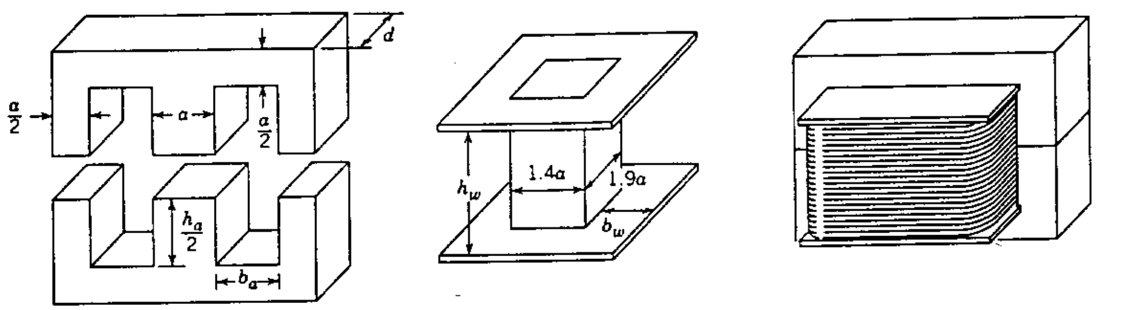
$J_{rms} \sim \frac{K_5}{\sqrt{K_{cu}}\ a}$. Likewise B_{ac} and core size are related as

$$B_{ac} \sim \frac{K_4}{f^{1/2}\ a^{2/5}}$$

$J * B \sim \frac{1}{a^{7/5}}$; and finally $LI(\max)\, I_{rms} \sim \frac{a^3}{f^{1/2}}$

The above relations show the importance of the chosen core size "a" in our quest for balancing J, B and inductor energy storage capability.

4. Typical Core Geometry's and Air Gaps For Inductors



All inductor cores have a puposeful air gap to store magnetic energy and to make the B-H curve more linear. The latter also allows for more inductor current to flow before B_{Sat} is reached as well as smaller cores to store a specific amount of energy. This "gapping of the core" does reduce the effective permeability of the core as the price for the more linear B-H characteristic. In an above "E-type core" the air gaps can be placed in one or more of the horizontal sections of the "E" structure. In general, three smaller air gaps are preferred to one big air gap because the flux leakage paths from one big gap extend further into wire winding window. This leakage flux will increase the proximity effect for wire losses in inductor wire windings in a local region causing non-uniform heating.

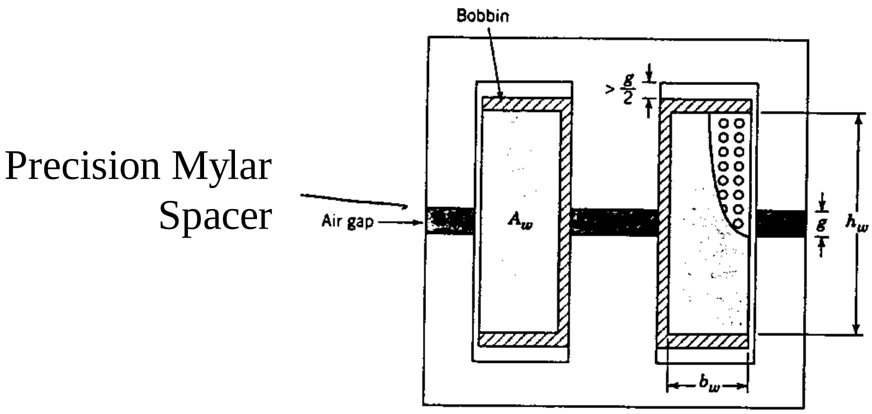


Figure 30-7 Cross-sectional view of a double-E core and bobbin assembled as an inductor. The air-gap spacing is supported by an insulating material (shaded area).

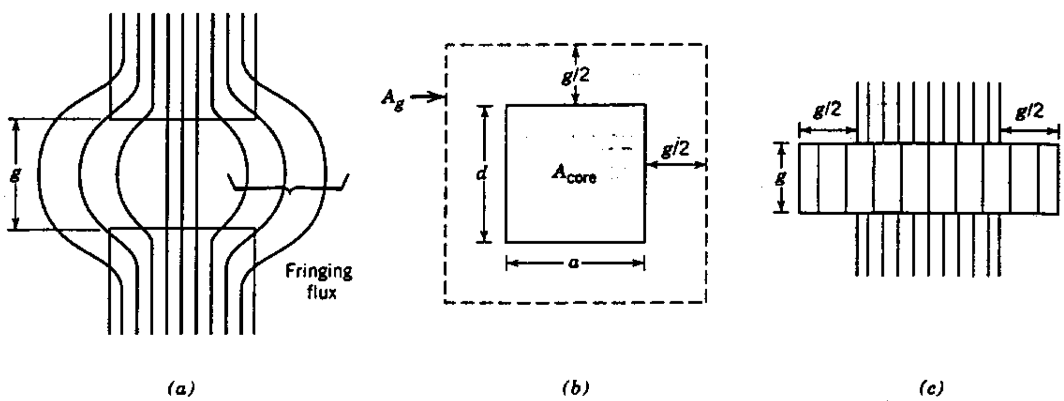
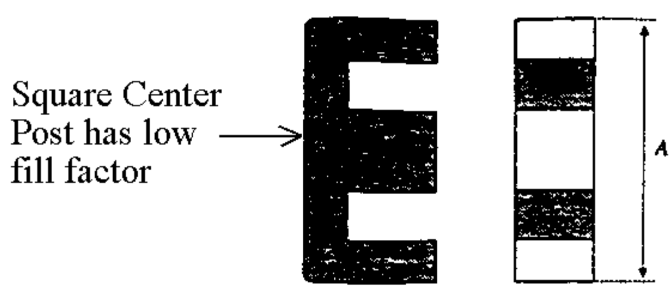


Figure 30-10 Fringing fields in the air gap of (*a*) an inductor (*b*) effective cross-sectional area of the gap and (*c*) an equivalent representation of the air gap.

Having talked about four general trends for cores we now need to get SPECIFIC regarding available commercial cores.

b. Specific Features in Selected Magnetic Cores

1. EE Core
Usually the lowest Cost Core and popular for profit margins



Circular posts have the minimum circumference and maximum core area, but square core cross-sections are available. Below the inductor transformer factor, K_g, is evaluated for a magnetic core loss exponent β = 2.7 for various EE cores. We give data for both K_g(inductor) and K_{gfc}(transformer) here to set the stage for transformer core choices as well. Note the large absolute difference between the two K factors for transformers versus inductors even though both scale with core size in a similar monotonic fashion as shown on the next page.

Can you tell whyfor HW #1 in general K_g(inductor)>K_{gfe}(transformer) ??? Bigger cores have bigger MLT for copper windings and weigh more as well as take up more physical room. On the other hand bigger cores have lower thermal resistance and can support more heat loss without reaching too high core temperatures. As a consequence POWER MAGNETICS is a field unto itself. We will only survey the highlights of this big topic.

EE CORES

Core type (A) (mm)	Geometrical constant K_g cm	Geometrical constant area K_{gfe} cm	Cross-sectional area A_c (cm^2)	Bobbin winding W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_e (cm)	Core weight (g)
EE12	$0.731 \cdot 10^{-3}$	$0.458 \cdot 10^{-3}$	0.14	0.085	2.28	2.7	2.34
EE16	$2.02 \cdot 10^{-3}$	$0.842 \cdot 10^{-3}$	0.19	0.190	3.40	3.45	3.29
EE19	$4.07 \cdot 10^{-3}$	$1.3 \cdot 10^{-3}$	0.23	0.284	3.69	3.94	4.83
EE22	$8.26 \cdot 10^{-3}$	$1.8 \cdot 10^{-3}$	0.41	0.196	3.99	3.96	8.81
EE30	$85.7 \cdot 10^{-3}$	$6.7 \cdot 10^{-3}$	1.09	0.476	6.60	5.57	32.4
EE40	0.209	$11.8 \cdot 10^{-3}$	1.27	1.10	8.50	7.70	50.3
EE50	0.909	$28.4 \cdot 10^{-3}$	2.26	1.78	10.0	9.58	116
EE60	1.38	$36.4 \cdot 10^{-3}$	2.47	2.89	12.8	11.0	135
EE70/68/19	5.06	$75.9 \cdot 10^{-3}$	3.24	6.75	14.0	18.0	280

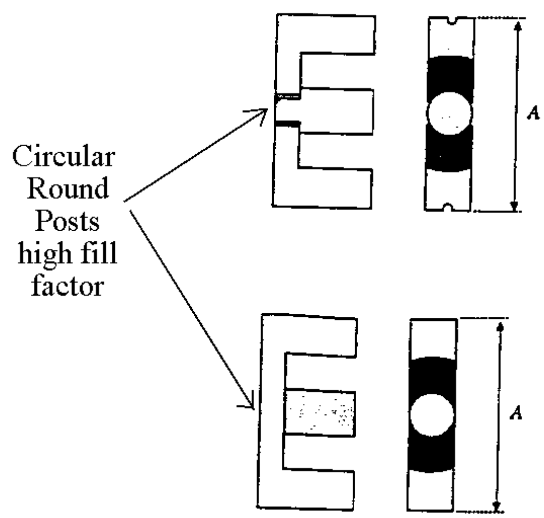
Note that large K_g and K_{gfe} values cost more money as the core size is bigger reequiring more ferrite material and bigger core molds. Large core size means lower B since $B = \phi/A_c$

2. EC Core

and very similar

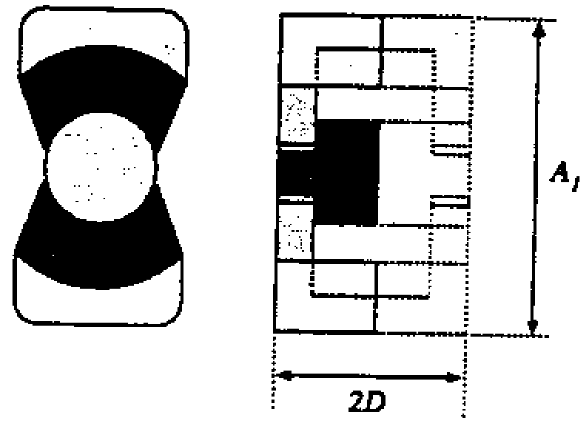
ETD Core

These cores usually employ core materials with the lowest thermal resistance, R_{th}, for use in hotspots.



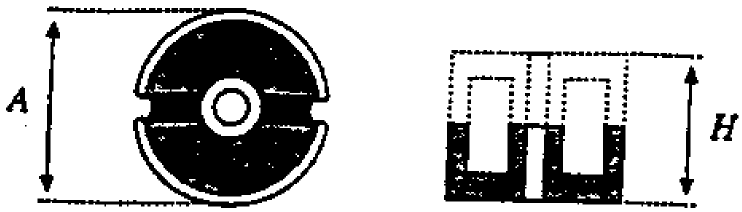
3. PQ cores
RN (not in appendix)

These cores have a minimum footprint or surface area and are usually made for printed circuit boards.



Core type (A$_1$/2D) (mm)	Geometrical constant K_g cm^5	Geometrical constant K_{gfc} cm^x	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm^2)	Magnetic path length l_e (cm^2)	Core weight (g)
PD 20/16	$22.4 \cdot 10^{-3}$	$3.7 \cdot 10^{-3}$	0.62	0.256	4.4	3.74	13
PQ 20/20	$33.6 \cdot 10^{-3}$	$4.8 \cdot 10^{-3}$	0.62	0.384	4.4	4.54	15
PQ 26/20	$83.9 \cdot 10^{-3}$	$7.2 \cdot 10^{-3}$	1.19	0.333	5.62	4.63	31
PQ26/25	0.125	$9.4 \cdot 10^{-3}$	1.18	0.503	5.62	5.55	36
PQ 32/20	0.203	$11.7 \cdot 10^{-3}$	1.70	0.471	6.71	5.55	42
PQ 32/30	0.384	$18.6 \cdot 10^{-3}$	1.61	0.995	6.71	7.46	55
PQ 35/35	0.820	$30.4 \cdot 10^{-3}$	1.96	1.61	7.52	8.79	73
PQ 40/40	1.20	$39.1 \cdot 10^{-3}$	2.01	2.50	8.39	10.2	95

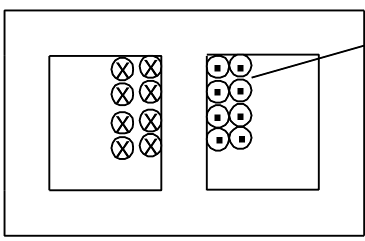
The large K_g and smaller K_{gfe} is clear trend for cores as is the trend to larger "K factors" for larger size cores. That is K_g(inductor) > K_{gfe}(transformer) because B_{max} of an inductor is higher due to a single winding with one directional MMF. In a transformer the mmf and net flux from the two windings act to cancel inside the core. Hence $K_{gfe} < K_g$. Finally, the issue of EMI is better addressed by closed path cores with the wire windings enclosed inside the core itself as show on the top of page 14 so that less leakage of flux occurs.



Closed Core or Donut Shape

Note the air windows for wire turns lie inside the core, where one or more Cu windings are wound. Flux leakage is minimized.

4. Pot Cores



By overlapping primary and secondary windings we achieve No/little leakage flux as the core shape surrounds the wire. Low leakage flux $\Rightarrow$ low L_l for transformer. Clearly as μ(core) increases we get less leakage flux. Still with $\mu = 5000$ nearly 0.2% of the magnetic flux goes into the air rather than in the core.

Pot Cores

Pot cores are specified by a four digit number (a b c d). Where the term a b is the core Diameter in mm and c d as the core height in mm. Again for transformers we prefer large c d values to achieve minimum H values. From Ampere's law as the l(air) will increase at fixed mmf, $NI = H \bullet dl$ and H may decrease thereby reducing stored energy outside the core by $[H(air)]^2$.

In the pot core table below the Geometric Constant K_{gfe} assumes $\beta = 2.7$. If the thermal resistance of the core(not shown in core tables) gets higher for a given power dissipation we get a hotter core surface temperature, T_s(core).

T_s must never exceed 100°C or the core material and perhaps wire insulation will degrade. For a fixed heat flow,Q, we know the thermal relation between the temperature difference of the surface and ambient temperatures, QR = ΔT. So smaller R allows bigger Q for fixed ΔT. Q arises both from Cu winding losses and from core losses. Both are heat sources making T_s rise. Below we include in the table of the core spec's the thermal resistance. Note the trend to lower thermal resistance with larger size cores.

Core type (A H) (mm)	Geometrical constant K_g (cm^5)	Geometrical constant K_{gfc} (cm^x)	Cross-sectional area A_c (cm^2)	Bobbin winding area W_A (cm^2)	Mean length per turn MLT (cm)	Magnetic path length l_e (cm)	Thermal resistance R_{in} ($^oC/W$)	Core weight (g)
704	$0.73 \cdot 10^{-6}$	$1.61 \cdot 10^{-6}$	0.070	$0.22 \cdot 10^{-3}$	1.46	1.0		0.5
905	$0.183 \cdot 10^{-3}$	256.10^{-6}	0.101	0.034	1.90	1.26		1.0
1107	$0.667 \cdot 10^{-3}$	$554 \cdot 10^{-6}$	0.167	0.055	2.30	1.55		1.8
1408	$2.107 \cdot 10^{3}$	$1.1 \cdot 10^{-3}$	0.251	0.097	2.90	2.00	100	3.2
1811	$9.45 \cdot 10^{-3}$	$2.6 \cdot 10^{-3}$	0.433	0.187	3.71	2.60	60	7.3
2213	$27.1 \cdot 10^{-3}$	$4.9 \cdot 10^{-3}$	0.635	0.297	4.42	3.15	38	13
2616	$69.1 \cdot 10^{-3}$	$8.2 \cdot 10^{-3}$	0.948	0.406	5.28	3.75	30	20
3019	0.180	$14.2 \cdot 10^{-3}$	1.38	0.587	6.20	4.50	23	34
3622	0.411	$21.7 \cdot 10^{-3}$	2.02	0.748	7.42	5.30	19	57
4229	1.15	$41.1 \cdot 10^{-3}$	2.66	1.40	8.60	6.81	13.5	104

For the above pot cores, employed with inductors, we find a K_g range:

$$\mathbf{10^{-7} \quad \underline{<} \quad K_g \quad \underline{<} \quad 1}$$

Now K_g has a term due to the core area (A_c) and two terms due to the wire windings area and wire length(W_A and MLT). Specifically the dependence of K_g is as shown below:

$$K_g = \frac{A_c^2 \ W_A}{MLT}$$

A_c^2 term corresponds to using more or less "iron or ferrite" in the magnetic device and the term W_A/MLT corresponds to using more of less "copper". For <u>fixed K_g</u> in a specific inductor core design we can trade Cu winding factors for Fe core factors.

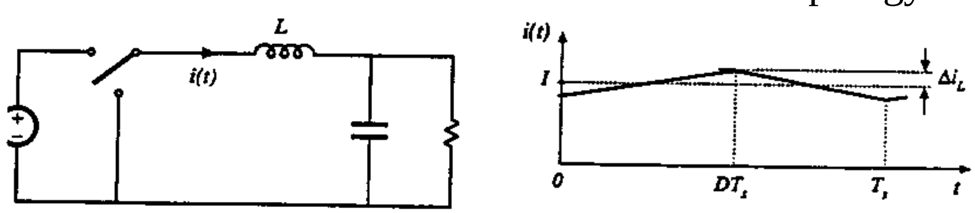
$A_c \uparrow$ gives a larger size core with lower thermal resistance, $\Re$(thermal). Larger core size also means Bmax is reduced so we have less core heating. MLT $\uparrow$ means we have higher resistivity per turn, which drives I^2R losses up. What core choices one makes depends on the desired circuit use we make of inductor as discussed next.

B. Types of Inductors

The design of the inductor windings and core parameters depends on the expected I_L(waveform) and the required inductance values as well as the operating frequency of the inductor current. Four examples of inductor use include: filter inductors on both the input and output of PWM converters, AC inductors in resonant converters, the two winding inductor with an air gap employed in a flyback converter, and a magnetically coupled inductor pair used in multiple output forward converters. All are discussed below

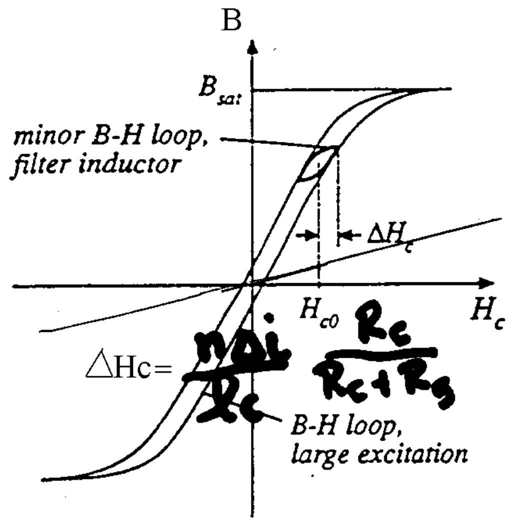
1. Filter Inductor/Coupled Inductor

a. Filter Inductor L in A Buck Circuit Topology



$I_{ac} << I_{dc}$

Here, core saturation of the filter inductor, which carries a high DC current, is the major risk.. Copper winding losses are <u>small</u> since I_{ac} is small. The DC I^2R losses often dominate over the AC wire winding losses, $P_{CU} = I^2{}_{rms} R$. Small I_{ac} means on a big DC level also means the associated region of the B-H region of the core is small, so core losses are at a minimum. Here in FILTER inductor design, we may wish to sacrifice copper wire loss for increased core loss. For the high DC current level we choose a core material of as high B_{sat} as possible for our primary goal, so no danger of core saturation occurs. If $L \rightarrow 0$ unexpectedly this tends to kill switches due to big current spikes when i > i(critical).



If we choose, a lower B_{sat} material for the core, we would need a larger geometry core so as not to exceed B_{sat}. We can of course in the inductor design trade off core material choice (B_{sat}) versus the required core geometry.

For the air gap reluctance greater than the core reluctance $Ni = \phi$ R(air gap) and $NI_{MAX} = B_{MAX}$ A(core) R(air gap) $= B_{MAX}$ *l(air gap)*/ μ_0. Here the turns ratio and the air gap are unknown for specified I_{MAX} and B_{MAX}.

The specified inductance relates the turns ratio,N, the core area, A_C, and the air gap length, *l(air gap)* as follows:

$$L = N^2/R(\text{air gap}) = \mu_0 A_C N^2 / l(\text{air gap})$$

Another constraint on the filter inductor design is the resistance of the filter inductor which is given by:

$$R = \rho N(MLT)/A_W(\text{wire windings})$$

Where the length of the wire comprising the N turns is set by the mean length of the turn on the chosen core times the number of turns.

A final constraint is the fact that the total wire coil area, N A_W, must fit within the winding window of the core, W_A. **Hence, K_u W_A.> NA_W** Where K_u is the fraction of the core window that is filled with copper that depends on wire size, wire shape and insulation. Typical values of K_u vary from .05 for high voltage inductors to 0.9 for low voltage foil-winding inductors.

The four constraints:

$$nI_{max} = B_{max} \frac{l_g}{\mu_0} \qquad L = \frac{\mu_0 A_c n^2}{l_g}$$

$$K_u W_A \geq nA_W \qquad R = \rho \frac{n\,(MLT)}{A_W}$$

These equations involve the quantities

A_c, W_A, and MLT, which are functions of the core geometry,

I_{max}, B_{max} , μ_0, L, K_u, R, and ρ, which are given specifications or other known quantities, and

n, l_g, and A_W, which are unknowns.

Eliminate the three unknowns, leading to a single equation involving the remaining quantities.

Elimination of n, l_g, and A_W leads to

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

- Right-hand side: specifications or other known quantities
- Left-hand side: function of only core geometry

So we must choose a core whose geometry satisfies the above equation.

The core geometrical constant K_g is defined as

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

We will summarize below the way we can achieve smaller and higher K_g values for cores employed as inductors. We do so in terms of the value of the required inductance ,L , the maximum inductor current expected ,I_{max}, the maximum allowable B in the core of the chosen material and the wire resistively as shown below.

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

K_g is a figure-of-merit that describes the effective electrical size of magnetic cores, in applications where the following quantities are specified:

- Copper loss
- Maximum flux density

How specifications affect the core size:

A smaller core can be used by increasing

$B_{max} \Rightarrow$ use core material having higher B_{sat}

$R \Rightarrow$ allow more copper loss

How the core geometry affects electrical capabilities:

A larger K_g can be obtained by increase of

$A_c \Rightarrow$ more iron core material, or

$W_A \Rightarrow$ larger window and more copper

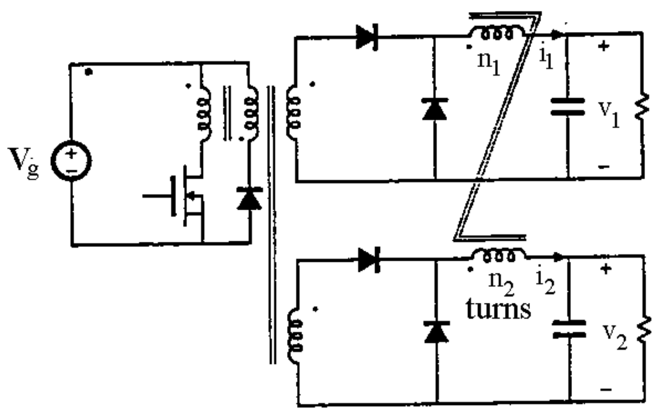
In summary, the core choice effects both magnetic and electrical properties of the inductor. In lecture 33 we will actually design an inductor making a variety of trade-offs. For now we will merely analyze the relationship K_g and explore trends with the various core parameters that we may specify.

b. Mutually Coupled Inductors

Windings n_1 and n_2 are on the same core but form two seperate but coupled inductors L_1 and L_2 as shown on page 18 for the two coupled circuits. **FOR HW #1 explain the reason you would employ mutually coupled inductors in the**

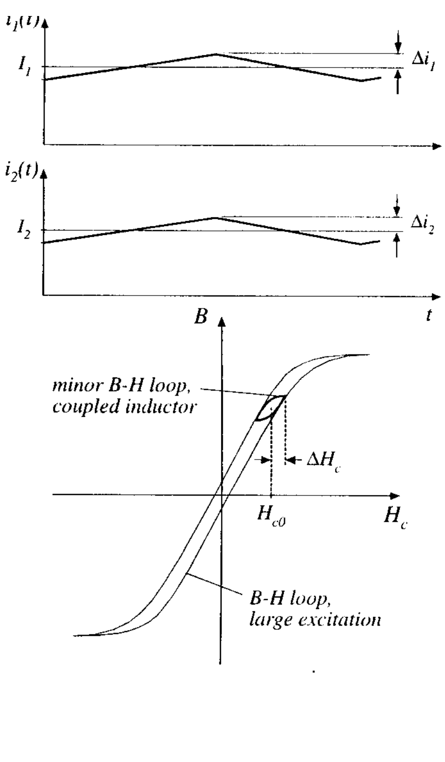
Two output forward converter circuit of page 18.

As a second example of coupled inductors revisit the input EMI filter we covered first semester and explain why coupled inductors do a better job than uncoupled ones.

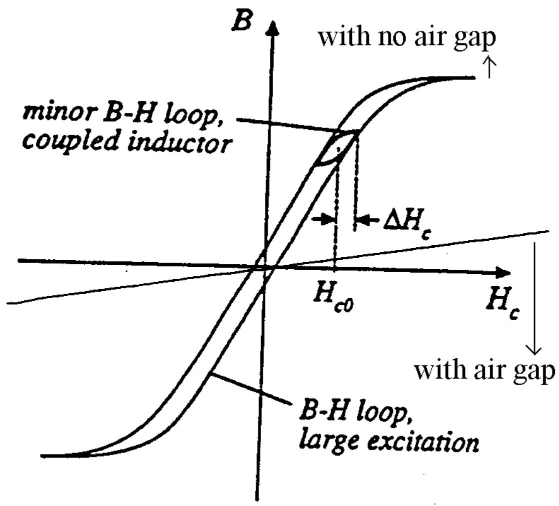


Again $I_{ac} << I_{dc}$ for proper operation L is controlled by tailoring the winding leakage inductances.

Below we plot the current waveforms and B-H region of operation



As discussed previously, we can add an air gap to the inductor core to reduce H(core) and thereby also the H(air window). H(window) is μ_c/μ_0 times H(core). Note the change in the B-H curve.



For coupled "L" on one core with additive mmf windings

$$H_c = \frac{N_1 i_1 + N_2 i_2}{\ell_c}$$

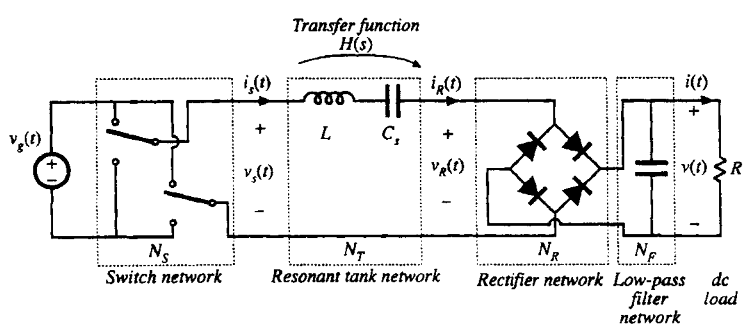
<u>Precision</u> mylar <u>sheets</u> can set l_g very precisely. As I_{ac} is so small compared to I_{dc} that we don't worry about it, the major wire winding loss is $I^2_{rms} R_L$,**with no proximity worries since $I_{ac} << I_{dc}$.**

2. AC Inductor: $I_{AC} > I_{DC}$

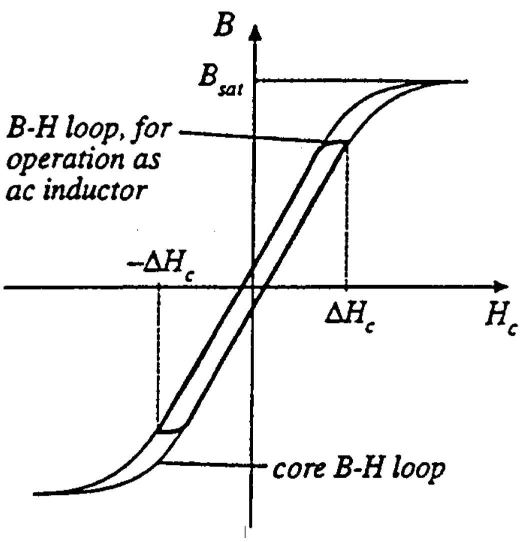
There are inductors whose I_L(waveform) is nearly pure AC. Below we consider several circuit topologies that satisfy this case. One will be the resonant converter inductor and the second the Flyback inductor. We will cover these inductors in Lecture 36 in detail.

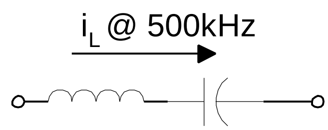
a. Resonant Converter(Erickson Chapter 19)

We are talking about the L in the resonant tank circuit below .

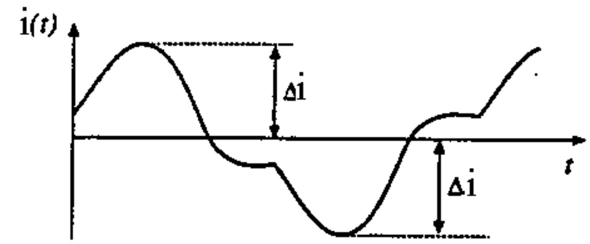


We employ a series resonance circuit tuned to f_{sw}, containing the inductor of interest, to extract from an applied square wave at f_{sw} only the fundamental sinewave component at f_{sw}. Here we employ a big B-H loop in the operation of the inductor. We will achieve a lower loss core if we can keep B_{max} low. Large size cores are always required since I_{ac}, H_{ac} and ϕ_{ac} are all large. We need big core area to reduce the expected B_{max}.
The i(t) corresponding B-H curves are shown below.





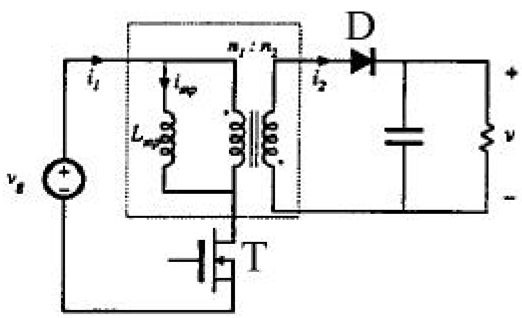
Usually to a first approximation ESR of "C" is negligible as is it for "L"



In resonant converter inductors $I^2_{rms}\ R_L$ (copper wire) and P(core) both dominate the total losses that will act to heat up the surface temperature of the chosen core. $P_v \sim K_{gfc}\ B^b_{max} \rightarrow$ Choose low loss ferrites which have low K_{gfe} (at f_{sw}).

b. The two winding core with an airgap employed in a Flyback Converter

In a flyback circuit, the air gap in the inductor core stores the energy placed there by the primary current winding during the first half-cycle. During the second half-cycle the stored energy in the inductor is transferred to the secondary winding.

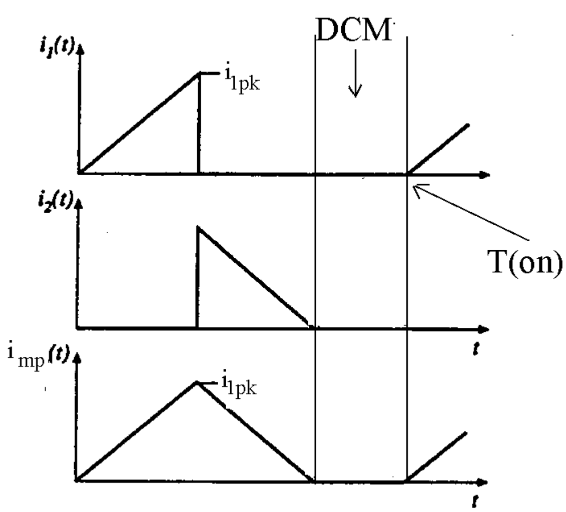


Summary of Operation

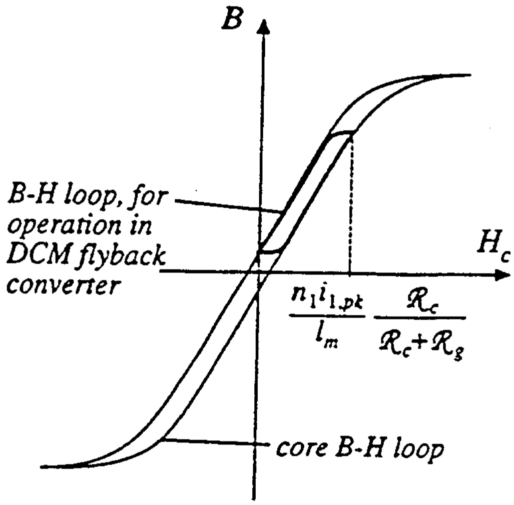
<u>T(ON) and Diode off</u>: i_{mp} flows through L_{mp} but there is no current in coils n_1 or n_2 of the ideal transformer.

<u>T(OFF) and diode on</u>: i_m flows <u>out</u> of coil n_1 from dotted side.
i_2 flows <u>into</u> coil n_2 on the dotted side and turns on the Diode delivering energy to the load.

Below we consider the DCM of operation of the flyback that avoids left hand plane (LHP) zeros in the AC transfer function that we will derive in later lectures. Note also $I_{AC} >> I_{DC}$ for DCM.



The DCM current waveforms will set the inductor currents and therefore the B-H utilization curves of the core employed in a FLYBACK converter. In particular the current is unipolar and goes to zero each switch cycle so only one quadrant of the B-H curve is employed.

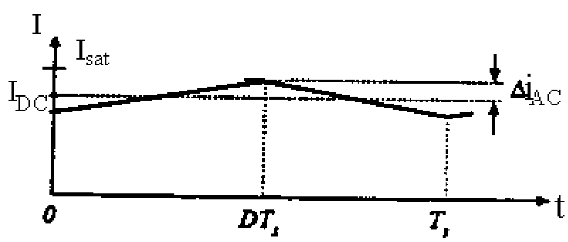


DCM mode H starts at “0"each switch cycle when T(ON), guaranteeing no core saturation. We use only 1/2 the core B-H loop as i_m is <u>unipolar</u>. With an airgap only a fraction of the H field appears across the core

Major power loss is core loss, if the wire proximity effect is avoided. Stored energy is crucial to flyback inductors.
$LI(max) \equiv \phi(max)\, N$ but because of complex inductor current waveforms $LI(max)I_{rms}$ is a key quantity to employ for better understanding of inductor requirements.

The Flyback inductance L_m is made purposefully small so that i_m is large. This i_m rise versus time is linear going as V_{in}/L. An air gap is used to store energy from the T(ON) cycle and deliver it to coil n_2 when T is off and the secondary diode conducts.

In CCM operation ac loss is usually low in core and in this flyback inductor $I_{DC} > I_{AC}$ as shown below in the current waveform.



Danger is I_{peak} exceeds I_{sat} for core for CCM operation and a circuit catastrophe occurs when L is shorted.