LECTURE 25

Basic Magnetic Material Information and Relation to Electrical Properties

Smaller size, lighter weight, lower loss and higher frequency magnetic components are the primary goals we are seeking.

A. Overview
   1. Inductors (high stored energy)
   2. Transformers (high transferred energy)

B. Magnetic Field Fundamentals

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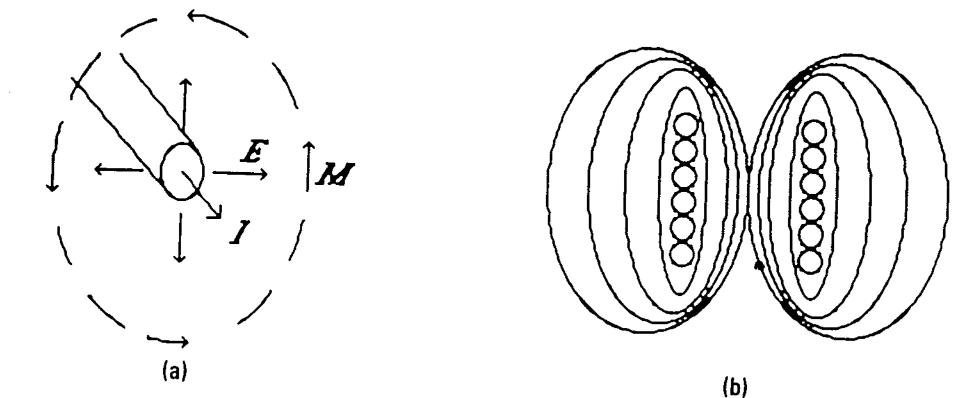
LECTURE 25

Basic Magnetic Material Information and Relation to Electrical Properties

A. Overview

Magnetic devices form the throbbing heart of switch mode power supplies. The inductors are the electrical flywheels smoothing out pulsed waveforms and the transformers make for easy voltage transformations. Both passive magnetic devices involve only copper wires and magnetic cores. The copper wire will carry a time varying current. The time varying current flowing in a single wire will cause an encircling magnetic field around the single wire, as shown below in Figure a. This allows for wires to interact at a distance and with proper design to form either inductors or transformers.



The fields around a conductor in free air: (a) field around a free wire; (b) field around a coil in free air.

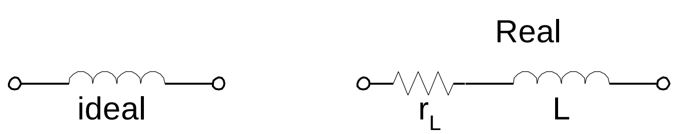
If an extended piece of wire is physically wrapped around into a stack of turns, the magnetic fields in free air will encircle all the wires as shown in Figure b above with an extended penetration of the field into space. This extended field distribution limits both the inductance's we can achieve in a small area footprint as well as the ability to make a good transformer as we will see herein.

We will employ magnetic cores to concentrate the magnetic fields in the regions around wires as described herein to achieve higher value inductors of smaller size. This concentration of magnetic field also allows a second copper coil to be wound such that the entire magnetic field from the first coil couples to the second coil. This tight magnetic coupling will allow for the design of a transformer with very little energy storage and efficient energy transfer between coils as detailed in the lecture.

The time varying magnetic field in the core itself will lead to core losses and heat generation. Surprisingly in addition, the magnetic field outside the core region that passes through Cu wires carrying current will also increase wire loss via additional **proximity effects**. That is the wire current will tend to further flow to the outside of the wire, as compared to that from simple skin effects, and cause increased resistance at high frequencies above the skin effect values by factors of 10-1000.

Below we give a brief overview of inductor and transformer characteristics as related to both the magnetic core and wire turns properties. Normal I^2R losses will not be sufficient for estimating wire losses of high frequency currents due to skin effects inside the wire which force high frequency current to flow on the outside surface of the wire and not flow at all in the center of the wire.

1. Inductors: A single Cu wire wound around a magnetic core. **The purpose of an inductor is to store electrical energy.** Storage will best be done in air, not in magnetic material as we show below. Hence, inductors have air gaps purposefully placed in their cores



Ideal Inductor	Real Inductor
1. L constant $L \neq f(w)$	1. L is a function of frequency via $\mu(f)$ of core. At High f, L decreases.

2. $r_L = 0$

2. $r_L \neq 0$ and $r_L(f)$
r_L increases with frequency due to both core and wire losses
   a. Core losses
      1. hysteresis core loss
      2. eddy current core loss
   b. Wire losses
      1. skin effect in one wire
      2. proximity effects in many layers of wires

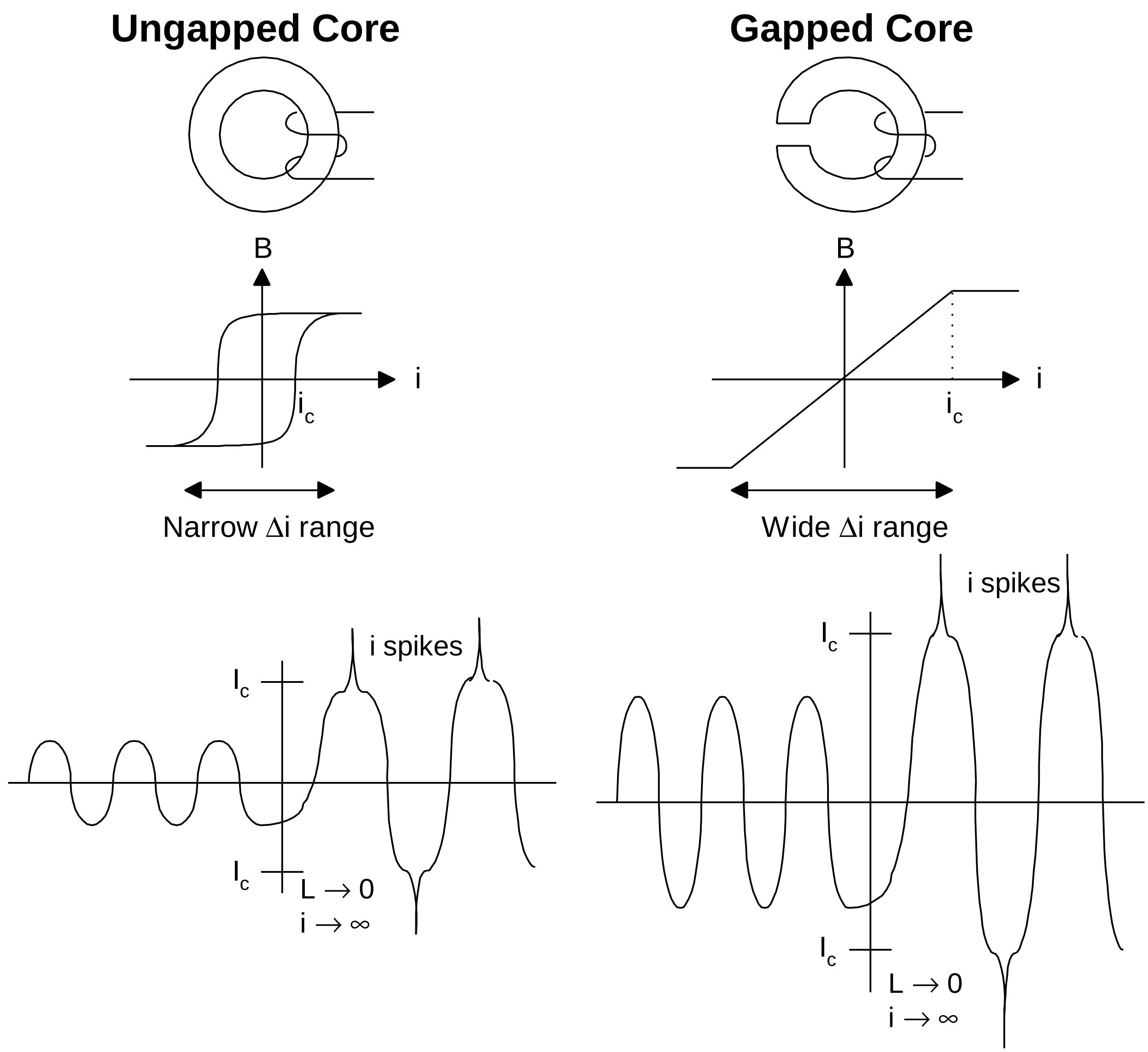
3. $L \neq f(i)$

3. L will decrease suddenly at high i due to **core saturation**. There exist a critical current level we cannot exceed or the core will saturate and $L \rightarrow 0$.

$$\varepsilon_{core} = \frac{1}{2m} H_c^2 \quad w / cm^3 \qquad \varepsilon_{gap} = \frac{1}{2m_o} H_g^2 \quad w / cm^3$$

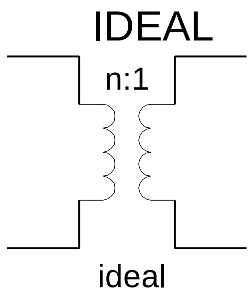
In summary, for $H_c = H_g$ we find the energy storage in the core ε_{core} is much LESS than the energy stored in the air ε_{gap} since the permeability of the core is 10-1000 that of air. That is air gaps will store more energy than magnetic materials. Since the purpose of inductors is to store energy, any core used on an inductor will have a gap cut in it.

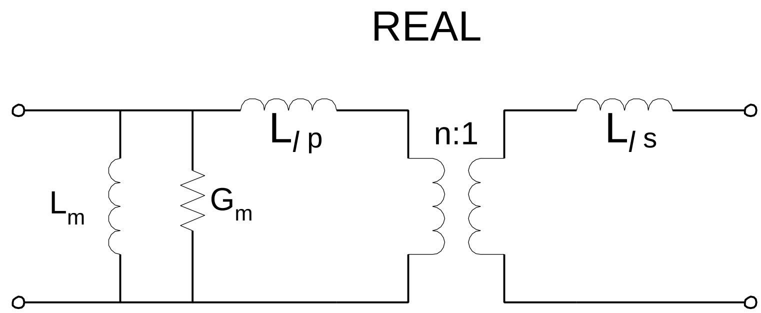
The figures below show another feature of cores with an air gap cut into them- improved saturation characteristics. The great advantage of cores is smaller more compact inductors. But the risk introduced to shrink geometrical size is that of saturation of the core at some value of current. This causes L to suddenly decrease. Lower or even shorted values of L will cause circuit problems in converters, like solid state switch failure. The current waveform will act normally until suddenly spikes of current appear as shown on page 5. Air core inductors have no saturation effects as they do not employ magnetic materials to concentrate flux.



We will show below in more detail how the particular non-linear B-H curves of a particular core material will cause differing hard or soft saturation effects for both inductors and transformers. The form of core saturation will effect the switch stress we create.

2. Transformers: Two electrically isolated sets of Cu wire coils wound around the same magnetic core. The **purpose of a transformer is to transfer energy** from one copper coil to another copper coil via the common extended magnetic field that couples both coils together. A well designed transformer stores no energy and this energy transfer role distinguishes it's role from that of an inductor. Coupling between coils is via the invisible to the eye B field. Tight B field coupling benefits from core material being present allowing a low reluctance path for field lines.





$L_{l\,p}$ is the leakage inductance due to finite permeability

1. Usually L_m is large in a well designed transformer but still draws some current which is negligable compared to the load current reflected into the primary. If the magnetic core <u>saturates</u> $L_m \rightarrow 0$, and we short out the ideal transformer. A flyback transformer is unique in that L_m is purposefully made small so I_m is large.

2. G_m loss due to
   core losses: both hysteresis and eddy current
   wire losses: both single wire skin effect and multi-wire proximity effects.

3. Leakage inductance L_l introduces series voltage drops and possible V transients due to inductive kick, when transformer currents are interrupted. Leakage flux arises because μ(core) only 5000 μ_o(air) therefore 0.02% of flux travels in the air not in the core. This creates leakage inductance of both the primary and secondary coils as we will see below. Usually, L_l is 1/50 of L_m due to finite μ(core).

4. Parasitic C <u>reduces</u> transformer high frequency response. C(parasitic) is mostly across the primary or secondary, but also occurs to a lesser extent between the primary and secondary coils which look to each other as opposing plates with slots in them.

The origin of all the above described issues for both inductors and transformers will be explained in detail below.

B. Magnetic Field Fundamentals

1. The Relation between B(flux density),H(field strength) MMF(magnetmotive force)and f(flux)

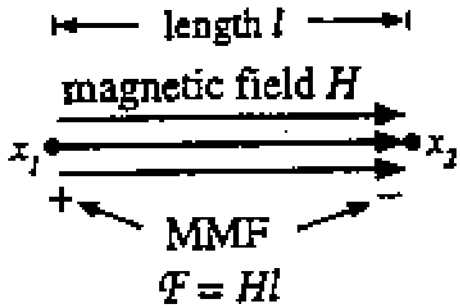
We need to define and distinguish magnetic Field intensity H, Magnetomotive force MMF = NI = F and magnetic flux ϕ vs. magnetic flux density = ϕ/A = B

a. Magnetic field intensity H is in units of A/M or A-turn/meter.

$$\oint H \cdot dl = \int_s \vec{J} \cdot \vec{S} + \frac{d}{dt}\int e\overline{E} \cdot d\overline{S}$$

This will form the x-axis in all B-H plots of magnetic materials. In power electronics the conductive current is typically J = 10^6 A/m^2 and it dominates over the displacement current, except at very high frequencies. We usually have E < 10^3 V/cm and the radian frequency of the currents is w = 10^7 rad/sec. Now for ε_o = 10 pF/m = 0.1 pF/cm, we find $w\varepsilon_o E$, the displacement current, is less than 0.1 A/m^2 or 7 orders less than the conductive currents and we can often say $\oint H \cdot dl = \int_s d \cdot dS = I$ (conduction).

Along a magnetic path of length l we find the MMF(magnomotive force) placed across the length l is related to the field strength H as shown below.



b. Magnetic Flux Density B: $B = m H = \frac{A\text{- }H}{m^2}$ $\; m$ in $\frac{\text{Henry}}{m}$

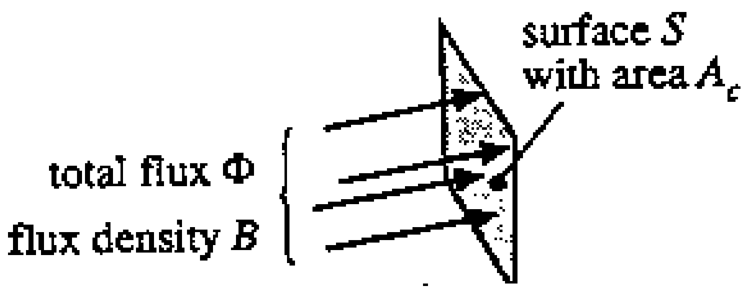
The realtionship between the B and H units is a complex one. For now, B is the magnetic flux density measured in Gauss or Webers per square meter. It will form the y-axis of all B-H plots for magnetic materials. The constant relating B and H is called the permeability of the medium with units of Henry/meter. We will see that the higher the permeability of the core the more inductance per turn of wire wrapped around the core. For air we find

$m_o = 4p*10^{-7}$ H / m $\approx$ 120 nH / m $\approx$ 1.2 nH / cm

Flux = Ampere*Henry = Li = ϕ = Webers

$B = \frac{f}{A} = \frac{\text{flux}}{m^2}$ = Tesla

$f(\text{webers}) = \vec{B} \cdot \vec{A}$



c. MMF(magnetomotive force) = $\int H \cdot dl$ = Ni

The closest electrical equivalent to mmf is voltage. The driving force to develop a magnetic field,H, with N windings of wire

carrying a current I will be given as shown below:
Where l_m is the length of magnetic path over which H extends.

$$H = \frac{NI}{l_m}$$

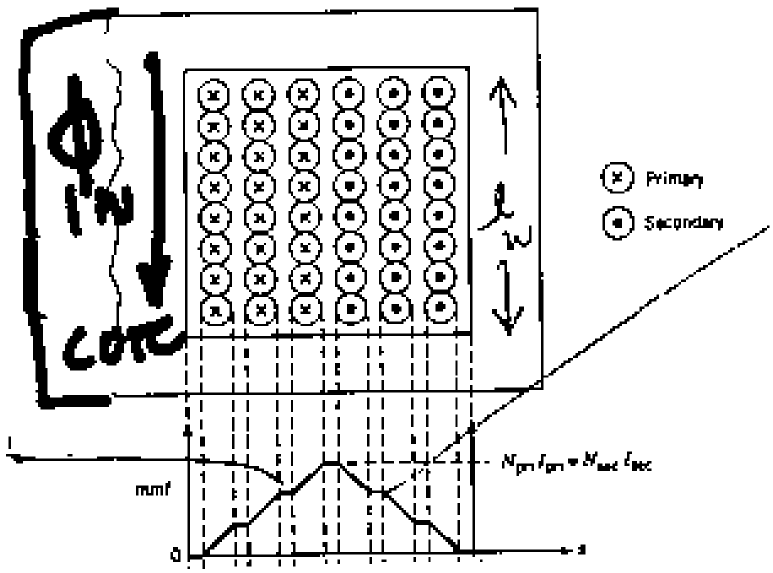
MMF is applied to a core of magnetic material by the product of the number of turns of wire and the current in the wire.
F = N*I = Ampere Turns (units of A-T)

$$\text{MMF} = \quad N_i I \;\equiv\; \underset{\text{magnetic field}}{H} \cdot \underset{\text{magnetic path}}{dl} \;=\; dF(\text{MMF})$$

Around a closed magnetic path $\int H \cdot dl$ = F (units of A-T)

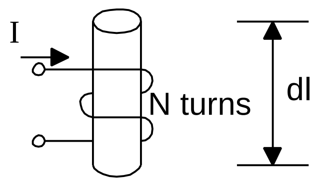
Note if N turns are wound in two layers around a core of total length d*l*/2 , rather than one layer of turns of twice the length, H is twice as large in the core. If N turns are wound in 10 layers of length d*l*/10 then H(core)is 10 times as much. The mmf inside the coil increases by using stacked layers of turns compared to one layer of turns. To further illustrate this point, below we consider the magnetic field that exists in the air window region of a open core. It is here in the window that wires are actually wound around the core. What we are plotting below is the leakage mmf from the core that is seen by each layer of wires. Note that net mmf increases for layers of coils carrying the same currents according to NI rules and then decreases as we encounter layers of coils that are carrying opposite directed current. **If we alternated coils carrying opposite currents would the mmf reach as high value**s??? We will find later that the resistive losses in the wires in various layers of turns will vary with the mmf that the wires are placed in or subjected to. That is the Ohmic resistance of wires wound with all primary turns wound first, allowing large MMF to exist, will differ from the resistance of INTERLEAVED turns of primary-secondary by factors of as much as 100 times. This is due to magnetic proximity effects which cause surface currents to be enhanced between nearby turns in proportion to the mmf seen by the individual wires. Details will be found in later lectures.

MMF plots in the winding window of a core versus distance from the center of the core are shown below. That is the NI cumulative effects are plotted versus wire position.



Note: $\vec{B}$ is parallel to wire windings.

We plot MMF which occurs to the core window.

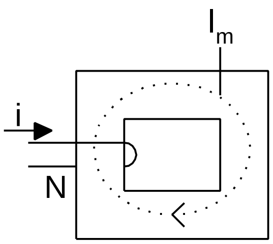


The relationship of NI= H x dl is termed Amperes Law and will be more fully covered in section 3 below.

Amperes Law states that the total magnemotive force F, along a closed path encircling the coil of wires carrying current is proportional to the ampere-turns in a winding. The magnetic field intensity is expressed in ampere-turns. When the field intensity H varies along the path, H must be integrated along the path length. Fortunately, the simplified form shown below can be used in most cases that H is constant along the path.

F = $\int H \cdot dl$ = NI = Hl in A-Turn units.

For an ungapped core, with closed magnetic path of length l_m, wound with N turns of copper wire carrying a current I we find:



$N*I = H(core) * l_m(core)$

Direction of H from the right hand rule; closed path

The magnetic material could have an air gap cut in the core. For a gapped core, wound with N turns carrying a current I we find there are two separate regions that drop mmf each with a unique H:

$$NI = H(core)\, l(core) + H(gap)\, l(gap)$$

2. System of Units for Magnetics

Unfortunately there are **<u>two systems of units</u>** for magnetic quantities the MKS system (Europe/Japan) and the unrationalized cgs. For the three main magnetic quantities we have a brief table showing this below.

magnetic quantity	MKs	unrationalized cgs	conversion Factors
core material equation	$B = \mu_o\mu_r H$	$B = \mu_r H$	
B	Tesla	Gauss	$1T = 10^4 G$
H	Ampere/meter	Oersted	$1\ A/m = 4\pi*10^{-3}$ Oe
Φ	Weber	Maxwell	$1Wb = 10^8$ Mx (New)
			$1T = 1Wb/m^2$

USA core manufacturers use the unrationalized units in the USA only, and unfortunately we have to be able to convert between these systems of units. Below we briefly describe the differences in SI and CGS units and then on page 12 give a full table of magnetic parameters with conversion factors. The internationally accepted SI system of units (Systeme international d'Unites) is a rationalized system, in which permeability, $\mu = \mu_o*\mu_r$ (μ_o is the absolute permeability of free space or nonmagnetic material = $4\pi*10^{-7}$; μ_r is the relative permeability of a magnetic material). In the unrationalized CGS system, $\mu_o = 1$, therefore μ_o is omitted from CGS equations so that $\mu = \mu_r$. But the rationalization constant μ_o doesn't just disappear in the CGS system - instead, portions of this constant show up in all the CGS equations, complicating them and making them more difficult to grasp. In the SI system, all of the “extra baggage of constants” is gathered into μ_o, thereby simplifying the SI equations.

The equations below are given in both systems - SI and CGS. It is suggested that beginners in magnetics design stick to the SI equations and ignore the CGS system until completely comfortable with the principles involved. Then, it may be helpful to use the CGS system when working with magnetics data expressed in CGS units, rather than convert the units.

SI and CGS Magnetic Parameters and Conversion Factors

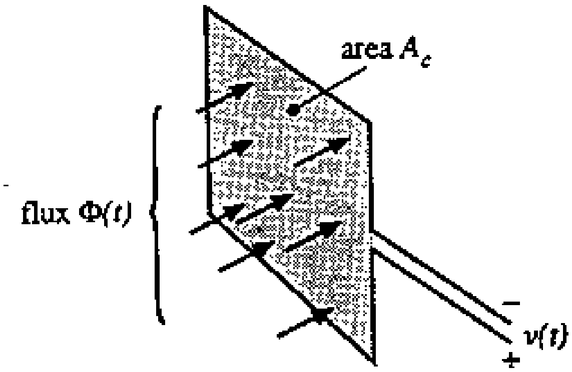
		SI	CGS	Conversion CGS to SI
Flux Density	B	Tesla	Gauss	10^{-4}
Field Intensity	H	A-T/m	Oersted	$1000/4\pi$
Permeability (space)	μ_o	$4\pi*10^{-7}$	1	$4\pi*10^{-7}$
Permeability (relative)	μ_r			1
Area(Core window)	A_e, A_g	m^2	cm^2	10^{-4}
Length(Core, Gap)	l_e, l_g	m	cm	10^{-2}
Total Flux = $\int BdA$	ϕ	Weber	Maxwell	10^{-8}
Total Field = $\oint H \cdot dl$	F, mmf	A-T	Gilbert	$10/4\pi$
Reluctance = F/ϕ	$\Re$	A-T/Wb	Gb/Mx	$10^9/4\pi$
Permeance = $1/\Re$	P	Wb/A-T	Mx/Gb	$4\pi*10^{-9}$
Inductance = $P*N^2$ (SI)	L	Henry	(Henry)	1
Energy	W	joule	Erg	10^{-7}

Above the magnetic reluctance, $\Re$, is analagous to electrical resistance and magnetic permeance,P, to electrical conductance.

3. Faraday and Amperes Law

a. Faraday’s Law: Open Loop of Wire in a time varying flux

A voltage is induced across the open winding(s). This corresponds to an electrical generator.



V(induced) = N $\frac{d f}{dt}$

= NA dB/dt
for harmonic time varying fields at w.
V = NA w B

Note polarity of induced V. Using the fact that the induced B field is opposing the B field applied we can see that V(induced) causes current flow such that current out of the wire coil is from the terminal labeled +. Check by right hand rule for + current flow to create an opposing B to the applied B. We will use either V or E to represent the induced voltage.

Faraday's Law
SI:

$$\frac{df}{dt} = -\frac{E}{N}; \qquad \Delta f = \frac{1}{N}\int Edt$$ Weber (N↑, ϕ↓)

CGS:

$$\frac{df}{dt} = -\frac{E}{N}x10^8; \qquad \Delta f = \frac{10^8}{N}\int Edt$$ Maxwell

Faraday's Law equates the flux rate of change, dϕ/dt, through a Cu wire winding to the volts/turn, -E/N, applied to the winding. Thus, the flux change is proportional to the integral volt-seconds per turn (directly equal in the SI system). Faraday's Law operates bilaterally - that is, if 2.5 volts/turn is applied to winding A, the flux through A will change by 2.5 Webers/second. If a second winding, B, is linked to all of the flux produced by winding A, then 2.5 Volts/turn will be induced in B.

Faraday's Law makes it clear that flux cannot change instantaneously. Any flux change requires a time to elapse and usually a change in energy. Time is required to move along the ϕ

or B axis, which is more obvious considering the electrical equivalent scale dimensions are Volt-seconds.

Note that all flux lines follow a closed loop path. Flux lines have no beginning or end- a kind of Indian life cycle philosophy. The energy stored in a magnetic field has lots of significance for inductors.

Magnetic Energy Storage

SI:

W/m^3 = $\int HdB \approx \frac{1}{2}BH = \frac{1}{2m}H^2$

W = $\int Vol \cdot HdB = \int I \cdot Edt$ Joules

CGS:

W = $\int Vol \cdot HdB = \int I \cdot Edt \cdot 10^{-7}$ Ergs

Energy put into and removed from the magnetic system can be determined by integrating the area between the characteristic B-H or ∫Edt-i and the vertical axis (B, ϕ, $\int Edt$) on the energy plane.

Energy must be integrated over time, which is a factor on the vertical axis, but not the horizontal.

b. General Energy Transformations Between Electrical and Magnetic Parameters

B-H magnetic curves can be transformed to ∫Edt vs I electrical curves as follows.

1. B $\rightarrow$ ∫Edt

B = ϕ*A

From Faraday's Law E = N$d\phi/dt$.

∫Edt = Nϕ

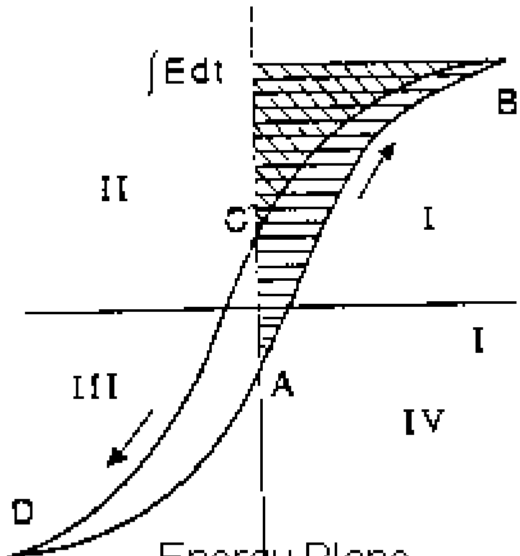
$\therefore$ B $\rightarrow$ ∫Edt

2. H $\rightarrow$ i

F = H * l

I = F/N

$\therefore$ H $\rightarrow$ I That is the x-axis corresponds to the wire current This magnetic to electrical transformation is shown in the figure below.



Energy Plane

It is much easier to understand this process by using the electrical equivalent axis, either Volt-seconds or Amperes. Referring to the energy plane above, from point A to B, energy from the external circuit is put into the magnetic system, as shown by the shaded area between A-B and the vertical axis. From B to C, magnetically stored energy is returned to the electrical circuit. The difference between the energy put in and taken out is hysteresis loss, the area between the two B-H or ∫Edt-i curves. At C, the magnetically stored energy is zero. From C to D, energy is put into the system. From D back to A energy is returned to the electrical circuit. The area between the curves is loss. At A, the remaining stored energy is zero.

+P $\Rightarrow$ Energy flow into magnetic circuit from the electric circuit.
-P $\Rightarrow$ Energy flow out of the magnetic circuit into electric circuit.

Positive energy sign indicates energy put in; a negative sign indicates energy returning to the external circuit. From A to B, voltage and current are both positive, so the energy sign is positive. Although the integrated Volt-seconds are negative at A, upward movement indicates positive voltage. From B to C, current is positive but voltage is negative (downward movement). Therefore the energy sign is negative. From C to D, current and voltage are both negative, hence positive energy. From D to A, negative current with positive voltage indicates returning energy.

c. Permeability, Permeance, Reluctance and Inductance

Permeability is a measure of - the amount of flux which a magnetic field can push through a unit volume of the magnetic material.

For free space $\mu_o = 4\pi*10^{-7}$ H/m or roughly 120 nH/m. This will change for a material medium according to the relations:

SI:

$\mu = \mu_o\mu_r = B/H$ Tesla/A-T/m

CGS:

$\mu = \mu_r = B/H$ Gauss/Oersted

When the material characteristic, permeability, is applied to a magnetic element of specific area and length, the result is permeance. **In the SI system, permeance is equal to the inductance of a single turn as we show below.**

Permeance, Reluctance

Permeance is roughly analogous to conductivity in the electrical realm. Reluctance = [Permeance]$^{-1}$.

SI:

$P = 1/\Re = \phi / F = BA / Hl$

$P = 1/\Re = \mu_o\mu_r A / l$ Webers/A-T

CGS:

$P = 1/\Re = \phi / F = BA / Hl$

$P = 1/\Re = \mu_o \mu_r A / l$ Maxwell's/Gilbert

Reluctance, the reciprocal of permeance, is analogous to resistance in an electrical circuit. (Don't push this analogy too far - reluctance is an energy storage element, whereas resistance is a dissipative element) Reluctance and permeance can be defined for the entire magnetic device as seen from the electrical terminals, but it is most useful to define the reluctance of specific elements/ regions within the device. This enables the construction of a reluctance model - a magnetic circuit diagram - which sheds considerable light on the performance of the device and how to improve it. From the reluctance model, using a duality process, a magnetic device can be translated into its equivalent electrical circuit, including parasitic inductance's.

Inductance

$\phi\Re = Ni = mmf$

Taking the time derivative of both sides and noting that the flux linkage $N\phi = \lambda$ and $d\lambda/dt = Vin$

$$\Re(d\phi/dt) = N(di/dt)$$

$$(\Re/N)(d\phi/dt) = (\Re/N)V_{in} = Ndi/dt$$

$$V_{in} = (N^2/\Re)(di/dt) = Ldi/dt$$

SI:

$L = N^2 P = \mu_o \mu_r N^2 A / l$ Henrys

CGS:

$L = 4\pi \mu_r N^2 A * 10^{-9} / l$ Henrys

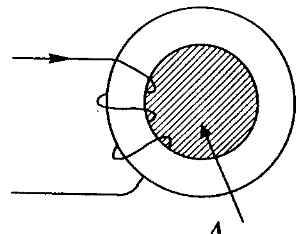
Inductance has the same value in the SI and CGS systems. In the SI system, inductance is simply the permeance times the number of turns squared. In all of the above equations the permeability of the magnetic medium is a design variable we can choose by specifying the core material. This effects stored energy as shown on page 18 below.

Role of μ_m in storing energy

For given values of L_m, I and B_{sat} ....

Core volume required $= \dfrac{L_m I^2 \mu_m}{B_{sat}}$

and $N^2 = L_m \dfrac{\ell_m}{\mu_m A_m}$



$\boxed{\text{As } \mu_m \uparrow\text{, required core volume} \uparrow, N \downarrow}$

N is related to required window area of the core

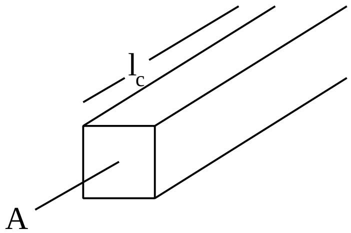
The cross section of wire is dictated by the current

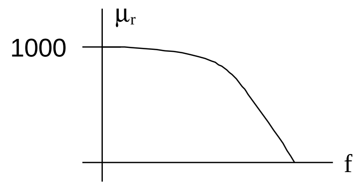
and $A_{window} \;\alpha\; N$

Role of μ_m in storing energy

- An appropriate μ_m is required for optimizing the design
- Two ways to achieve optimum μ_m
  - Distributed air gap as in powdered iron and MPP cores - where effective μ_r is few hundreds or less
  - Discrete air gap

In summary, Reluctance, $\mathfrak{R}$, of magnetic Materials [Henry^{-1} Units]





$$\mathfrak{R} = \frac{l_c}{mA_c} \; [H^{-1}]$$

where l(core) is in meters and μ is in H/m with core area A_c in m^2.

In a closed magnetic circuit as shown below $\mathfrak{R}$ is akin to resistance.

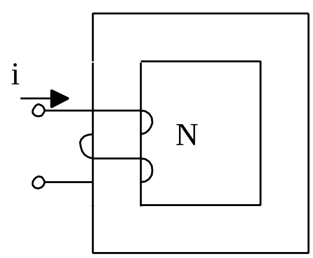
Permeance P = 1/$\mathfrak{R}$ is a concept akin to admittance of electrical circuits.

$$P = 1/\mathfrak{R} = \phi/F = BA/Hl = \mu_o\mu_r A/l.$$

SI units of Permeance are Webers/Amp-Turn.

In CGS: $P = 1/\mathfrak{R} = \mu_r A/l$ in Maxwells/Gilbert

d. Ampere's Law $\oint H \cdot dl$ = Ni



$Ni = F = mmf = \phi\mathfrak{R} = H_c l_c$

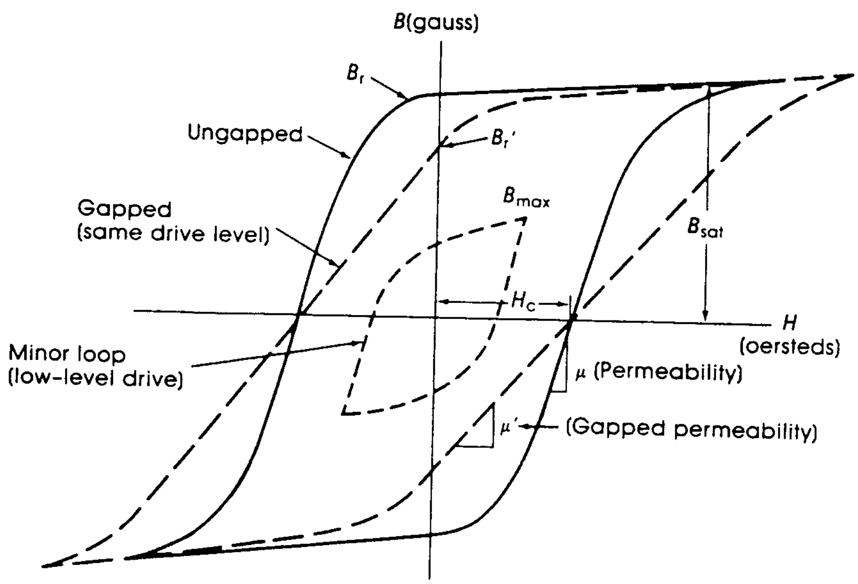
Ni is in A-turn and ϕ is in Wb with $\mathfrak{R}$ in H^{-1}.

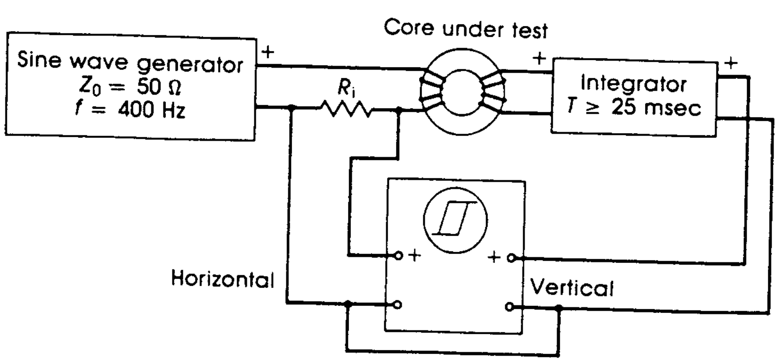
$L(H)i(A) = \phi(Wb)$

e. Measuring B-H Curves on Unknown Cores

Ampere's Law, Lenz's Law and Faraday's Law jointly govern the important relationship between the magnetic elements and the equivalent circuit as seen across the Cu wire windings. That is via these three laws we can make relationships between the electrical and magnetic circuits.

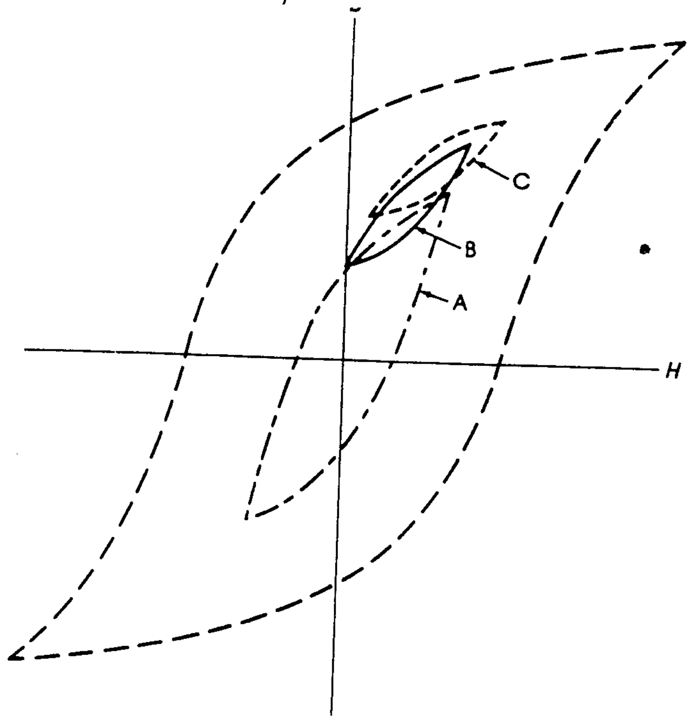
An unknown magnetic material could be characterized by measuring the B-H curves that relate the magnetic field intensity, H, to the magnetic flux density,B. This would be accomplished using the following experimental set-up. The applied current to the input coil of N turns would allow for the creation of a specific field intensity, provided we know the length of the magnetic path in the core under test, l_m. This information is fed into the horizontal axis of an oscilloscope as shown on page 12. The voltage induced on a second set of turns electrically isolated from the first coil is integrated to get a measure of the flux density,B. Below we show B-H curves for both gapped and ungapped cores and how this effects the saturation characteristics of the B-H curves. Note in particular the shift to higher H values for gapped cores.





The B–H curve and how to display it.

The operation region of the core depends on the current waveforms in the current carrying coils. The current waveforms depend on the location of the magnetic element in the electrical circuit. Below we show three operation regions for one core material whose B-H characteristics are shown by the outermost dotted lines. Inside this region we show three possible operating regions labeled: A,B and C.



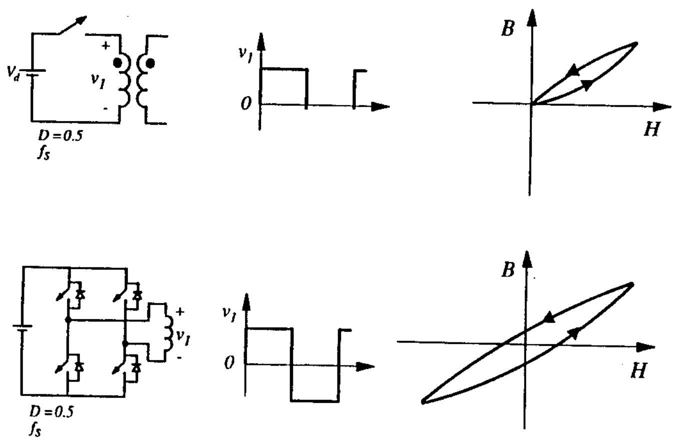
Minor loop B–H curves for various magnetic components.

Curve A represents a core operated in a bipolar manner such as the transformer core of a bridge or push-pull converter.
Curve B represents a core operated in a discontinuous mode flyback converter where the current goes from zero to positive but never goes negative.
Curve C represents the core used in a BUCK converter choke where there is a large DC level around which there is an AC ripple current, which never reaches zero current.

Can you think of further possibilities???

One can choose the switch topology in a switch mode power supply. We show below two different switch drive topologies that result in two different B-H utilization curves for the transformer in the two cases. Clearly, the cores in the two cases see two unique B-H conditions. If the size of the core is an issue this shows a way to reduce the core cross-section by a factor two. Use bipolar core excitation.

Core utilization in unidirectional and bi-directional magnetization topologies



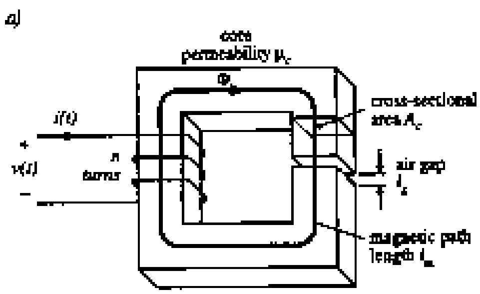
$$\Delta B = \frac{V_d \frac{T_s}{2}}{N_1 A_{C_1}} = B_{max}$$

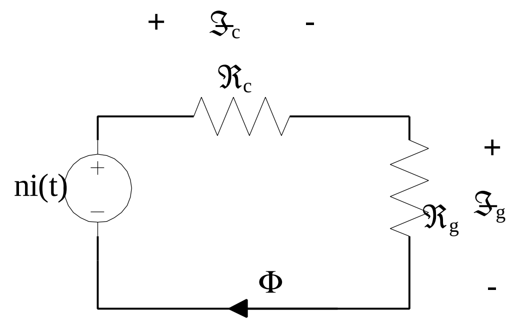
$$\Delta B = \frac{V_d \frac{T_s}{2}}{N_1 A_{C_2}} = 2B_{max}$$

$\therefore A_{C_2} = \frac{A_{C_1}}{2}$ (for the same no. of turns)
i.e., bidirectional core excitation results in smaller core volume

f. Air Gap in Magnetic Core Media

We review the solution to the magnetic circuits we face in inductor and transformer design on a DC basis. The bulk of the magnetic energy is primarily stored in the air gap not in the core itself.





Where for the core itself $\Re_c = \dfrac{lc}{m_c m_o Ac}$ and $\Re_g = \dfrac{lg}{m_o Ag}$. Due to flux fringing $A_g > A_c$. When a magnetomotive force F = ni is applied to the gapped core a flux flows. The voltage across the copper coils with n turns is:

$ni = \phi(\Re_c + \Re_g) = \phi\Re_T$

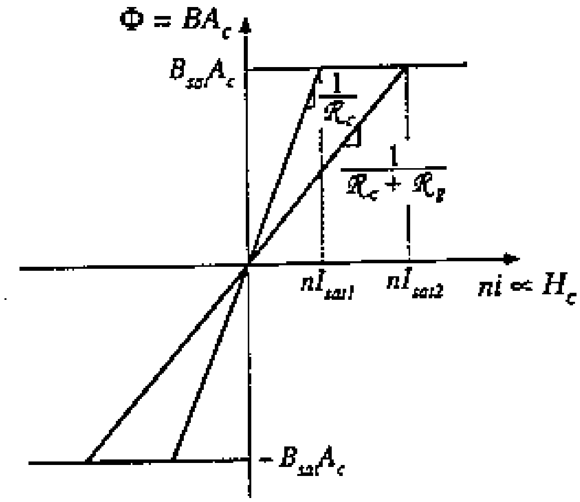
$$v = n\frac{d\,f}{dt} = L\frac{di}{dt} = \frac{n^2}{\Re_c + \Re_g}\frac{di}{dt}$$

Where the inductance $L \equiv \dfrac{N^2}{\Re_T}$ [H]

For higher magnetic reluctance values of the core upon which the wires are wound, we get lower inductance. Hence the airgap size in flyback transformer cores is big. For regular transformers no air gap is employed at all since energy transfer not energy storage is sought. Hence, the reluctance values of transformers wants to be as small as possible so that L is maximized.

The B-H curves vary as the gap size is increased making $\Re_g$ vary. For $\Re_g = 0$, $\Re_T = \Re_c$ but for large $\Re_g$, $\Re_T \sim \Re_g$. These two cases are shown below on page 24.

We can also create a distributed air gap by employing powdered magnetic materials that are coated with an inert electrical insulation to minimize eddy currents. This introduces a DISTRIBUTED air gap into the core structure. The insulated powder is compacted into a core geometry to achieve permeabilities of 10 to 1000.



Case #1: $\Re_g = 0$

$$\Re_C = \frac{l_C}{m_C m_O A_C}$$

$\mu_c = f(w,T,i)$

$\Rightarrow$ L(w,T,i) L is constant over small Δi range.

Case #2: $\Re_g >> \Re_c$

For $\Re_G > \Re_c$

$\Re_{total} \neq f(w,T,i)$

Since μ_o fixed.

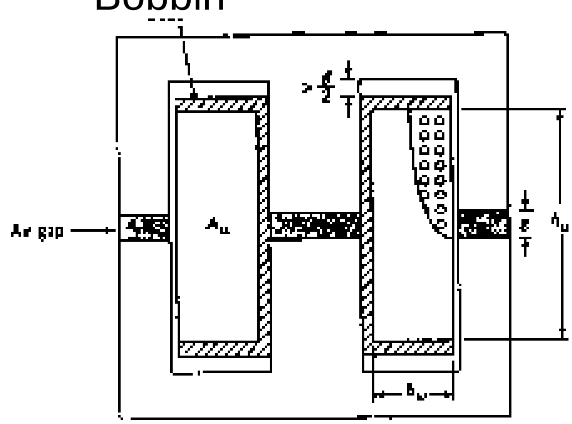
$\Rightarrow L \neq f(w,T,i)$

L(air gap) = constant over a larger Δi range.

Note above that I_{sat} with the air gap is larger than I_{sat} with no air gap but the absolute value of L over the Δi range is less!

For flyback converters the air gap region also acts as an energy storage element. Energy is stored in the air gap rather than the core because of the very large value of H(air) compared to H(core). Below we show a cross-section of an E-shaped core with wire windings on the center leg of the E structure. Note the two E shaped sections are separated by an air gap in three places

Bobbin



$$e = \frac{m_o}{2} H^2 * V(\text{volume})$$

$$H_{\text{air gap}} = \frac{m_c}{m_o} H_c$$

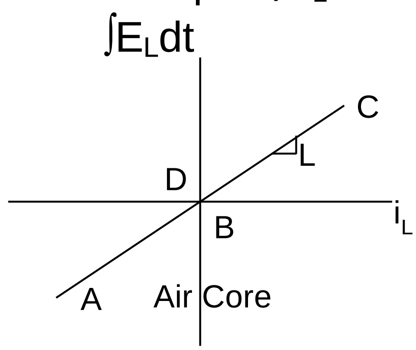
$\frac{m_c}{m_o}$ can be 10-10^5 so

ε(air gap) >> ε(core)

Next we compare a lossless air core inductor to an inductor made using a magnetic core.

g. Air Core Inductor versus Magnetic Core Inductor

Below we plot $\int E_L dt$ vs i_L

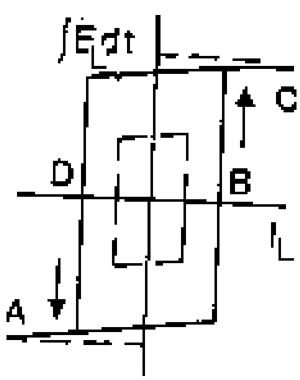


$(1/L) \int E_L dt = i_L$
i_L is in phase with $\int E_L dt$.
The slope is L^{-1}. For large L values the slope is small. For small L values the slope is large.

From B to C and D to A the product of voltage and current is positive and energy is put into the inductor.

Form C to D and A to B the product of voltage and current is negative and energy is removed from the inductor.

Below we plot $\int Edt$ versus iL for the magnetic core inductor where the core material has hysteresis.



(d)

The area within the characteristic is all hysteresis loss. The only energy returned tot he electrical circuit is the thin triangular wedges above and below the saturation values.

Only when the core saturates, acting like an air core, is any energy stored in the core returned to the electrical circuit. The interior minor loop occurs with reduced current drive where the core does <u>not</u> saturate.