

LECTURE 26

Losses in Magnetic Devices

Achieving acceptable loss levels in magnetic devices is the primary goal.

- A. Overview of Magnetic Device Losses
- B. Faraday's and Lenz's Law
 - 1. B_{\max} in Cores from Faradays Law
 - 2. Lenz Law and Its Implications for Induced Current Losses
 - a. Skin Effects in Copper Wires
 - b. Eddy Current Effects in Various Core Materials
- C. Magnetic Media B-H Curves and Losses
 - 1. Basics of B-H Curves
 - The Saturation Catastrophe
 - 2. Losses in Magnetic Media
 - a. Hysteresis Loss $\sim f(B_{AC})^m * \text{Core Volume}$
 - b. Eddy Current Loss $\sim f^2(B_{ac})^n * \text{Core Volume}$
 - c. High f Ferrite Loss $\sim f^3 * \text{Core Volume}$
 - d. Summary of Losses versus B and f

FOR HWS# 4 and 5 do the following:

HW#4 Erickson Chapter 12 problems 6 and 8 and in class questions as well as lecture questions

Hw# 5 Erickson Chapter 13 problem 3 and in class questions as well as lecture questions

LECTURE 26

Losses in Magnetic Devices

A. Overview of Magnetic Device Losses

1. General Situation

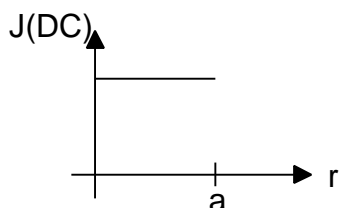
Magnetic devices are composed of copper wires and magnetic core material. Losses in magnetic devices occur both in the copper windings and in the magnetic core materials. One or the other loss mechanism can dominate depending on the particulars of the magnetic device operating conditions. **Properly designed magnetic devices** should add up to no more than **1-5 % of the total losses** in a switch mode electronics circuit. Solid state device losses will contribute a similar amount. Overall energy conversion efficiency should be 90 to 98 %.

2. Copper Wire Losses

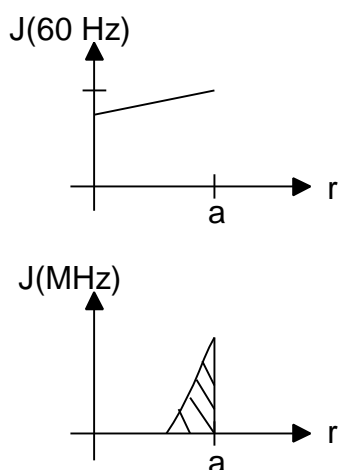
a. Skin Effects on an Isolated Single Wire

The wire losses are I^2R losses and the DC I^2R levels are a baseline for comparison to losses at any other operating frequency. The ISOLATED SINGLE wire resistance increases above DC levels by factors of 2-10 through **skin effects for high frequency currents**. The resistance of the wire is a function of the frequency of the applied current, f , in the wire and chosen the wire diameter, D . The larger the wire diameter the bigger the skin effect. **Roughly, $R_{AC}(\text{wire}) \sim R_{DC}(\text{wire}) f^{1/2} D$**

For now we intuitively plot the current density in a wire versus wire radius for the case of DC currents through some high frequency current to visually see the trends prior to detailed mathematical analysis.



For dc conditions J is uniform over the wire diameter as we expect.



Even at mains ac J (60 Hz) has a small dip in J near $r = 0$ of 1-3%. Why? This is well known to increase AC transmission line losses.

For MHz currents J flows only at or near the surface of the wire. No flow of current at $r = 0$. This makes wires look 2-100 times smaller.

We will see later in more detail how eddy currents in the copper wires induced by the time varying H field encircling the wire will alter J profiles as shown above to minimize total energy at the expense of increasing $I^2 R$ losses.

b. Proximity Effects on Multiple Turns of Wire

In multiple turns of wire as we find on all practical magnetic devices, each wire turn will have unique additional losses via **proximity effects due to net H fields from all nearby wires**. For proximity effects the cooperative effect of the net magnetic fields, generated by multiple turns of wire, changes the resistive power losses in a particular wire turn in a different way even from the nearby wire turns. Losses depend on the stray magnetic field from the magnetic device that passes through that individual wire turn. This field and associated MMF will depend on the wire turn's location in the device, usually within the air core window and also on the chosen winding configuration. As we will see below, proximity effects at AC frequencies alone can increase wire resistivity by factors of 10-1000.

3. Magnetic Core Losses

There will be two components to core losses. Both losses arise due to time varying magnetic fields inside the core. The physical mechanisms of the two core losses are quite unique even though both are caused by the same time varying magnetic fields. One is

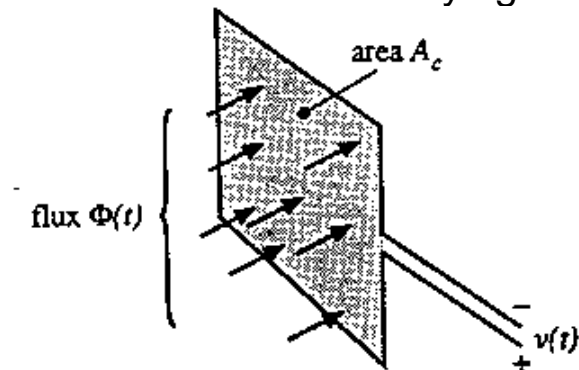
due to hysteresis loss in the B-H curve and it varies linearly with the frequency of the magnetic field variation. The second loss is due to eddy currents induced in the core material and it varies as the square of the frequency and inversely as the electrical resistivity of the core material.

All of the loss processes for magnetic devices are covered in detail in this lecture except proximity effects on resistivity changes for stacked turns of wire which are given in detail in lecture 28.

B. Faraday's and Lenz's Law

1. B_{\max} in Cores from Faradays Law

Faraday's law gave the voltage induced in an open loop of wire immersed in a time varying magnetic field as:



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

$$= NA \frac{dB}{dt}$$

for harmonic time varying fields at ω .

$$V = NA \omega B$$

We can see that $V(\text{induced})$ causes current flow such that current out of the wire coil is from the terminal labeled +. We can work this backwards to find the maximum B inside a core when we apply a voltage V across the coils that are wrapped around the core. This allows one to determine B_{\max} and compare it to the value of B_{sat} for the specific core material to avoid the saturation disaster when B_{\max} exceeds B_{sat} . From Faraday's law we find:

$$B_{\max} = \frac{V}{kNA} \frac{1}{f}$$

Where :

k is 4.44 for Sine waves ($2^{-1/2} \times 2\pi$) and 4 for squarewaves, etc.

N is the number of wire turns

A is the cross-sectional area of the core material

f is the applied frequency

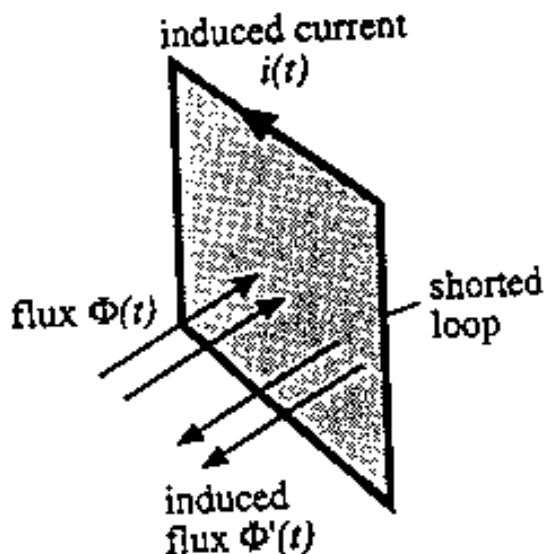
B_{\max} is also employed to find the hysteresis core losses which depend on peak B fields in the core rather than the RMS value of B. Design paths and compromises to lower B_{\max} values are clear.

2. Lenz's Law and Its Implications to Magnetic Device Losses

a. Current Loop

Placing a CLOSED loop of wire in a time varying magnetic field will induce a current in the wire to oppose the applied B field as shown below. This is the basis of electric motor design.

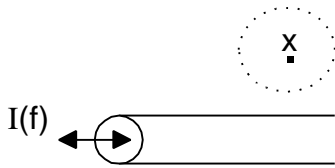
Lenz's Law: Loop of Shorted Wire in a $\phi(t)$ magnetic flux field
This corresponds to the situation in an electrical motor.



Induced i flow in the closed loop of wire at one moment of time is such as to oppose the applied flux at that same moment as seen by the right hand rule.

b. H Field Surrounding An Isolated Single Wire

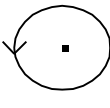
Next consider a single isolated wire with a sinusoidal current at a frequency f flowing through it. A sinusoidal current i creates a \pm circular magnetic H field enclosing it. The field is moving in the θ direction both within the wire and outside the wire. Note also that the induced magnetic field can not distinguish the solid wire from the air surrounding it. Why? Is the permeability of copper different from air??



Since the permeability of air is indistinguishable from copper:

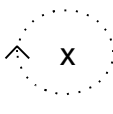
$$\mu(\text{wire}) = \mu_0(\text{air})$$

Current (out of page toward you) creates counterclockwise H field.

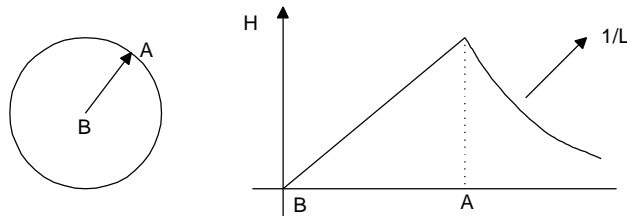
$$\oint \vec{H} \cdot d\vec{l} = I$$


From Ampere's Law at time $t = t_x$.

Current (into the page) creates a clockwise H field. The spatial variation of the H field with radial position for a constant current density versus wire radius is plotted below.

$$\oint \vec{H} \cdot d\vec{l} = -I$$


From Ampere's Law at time $t = t_y$.

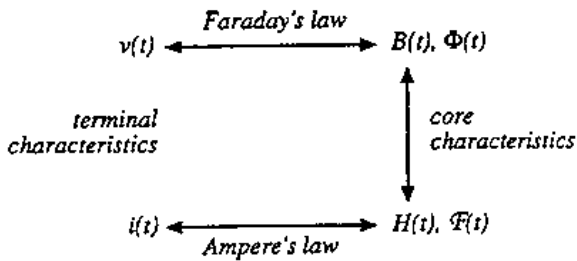


Does this above surrounding H field from a DC current around the wire have any effect on the current in the wire?? Note also that the H field penetrates into the wire itself. Stay tuned and you will see amazing interactions between the H field and the current density distributions when we make the current an AC rather than a DC one.

Magnetic effects on currents and current distributions, as we will unfold them, will explain many unusual phenomena encountered in power electronics both in wires and in core materials. For wires both skin and proximity effects have their root cause in H field effects. The eddy currents that are induced in core materials via Lenz's law cause heating of the core and loss of energy.

A overview of field and electrical interactions is on page 7.

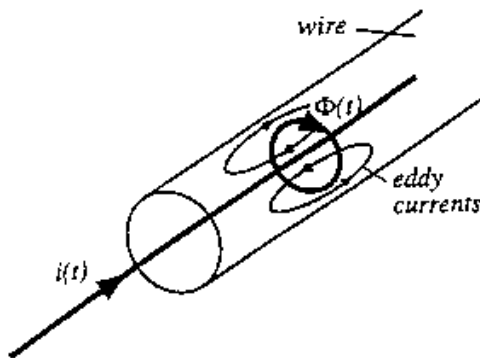
Circuits/Fields



Relations from circuit i, v to the materials B, H fields.

C. Skin Effects and Current Distributions Across Wire Cross-sections

The H fields surrounding the wire have little effect on the insulating air surrounding the wire. However, these same H fields when they penetrate the conducting wire itself will cause the following chain of events that alters the current density distribution in the wire from a uniform one versus radial position to a non-uniform distribution that pushes currents to the outer surface. One wire with ac current flowing causes a magnetic field enclosing the wire. In turn the time varying H field inside the wire causes additional current flow loops within the wire whose directions are revealed best by Amperes Law: $\int \mathbf{H} \cdot d\mathbf{l} = H2\pi r = I$.

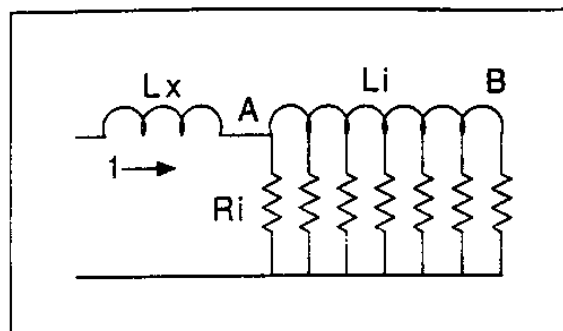


Flow of $i(t)$ causes $H_\theta(t)$ fields. $H_\theta(t)$ in turn causes elliptical voltages to appear in the wire which drive longitudinal elliptical eddy currents to oppose $H_\theta(t)$. The eddy currents act to cancel out the applied current in the center of the wire but ADD at the wire surface.

The induced eddy currents in the Cu wire act to enhance current flow at the edge of the wire but to reduce current flow at the center of the wire. Looking at a wire cross-section better explains the net effect on the $J(r)$ profile. Current flows in the path(s) that result in

the lowest expenditure of energy. At low frequency, this is simply accomplished by minimizing I^2R losses only. At high frequency however, current flow also occurs in path(s) that involve inductive energy. Now energy transfer to and from the magnetic field generated by the AC current flow, must also be minimized. Conservation of both resistive and inductive components causes high frequency current to flow nearer the surface of a large diameter wire conductor, even though this may result in much higher I^2R losses. If there are several available paths, HF current will take the path(s) that minimize inductive energy flow.

The circuit below shows an inductor in series with an L-R transmission line. What happens when a dc current is put through this circuit? Do you get a minimum resistance?

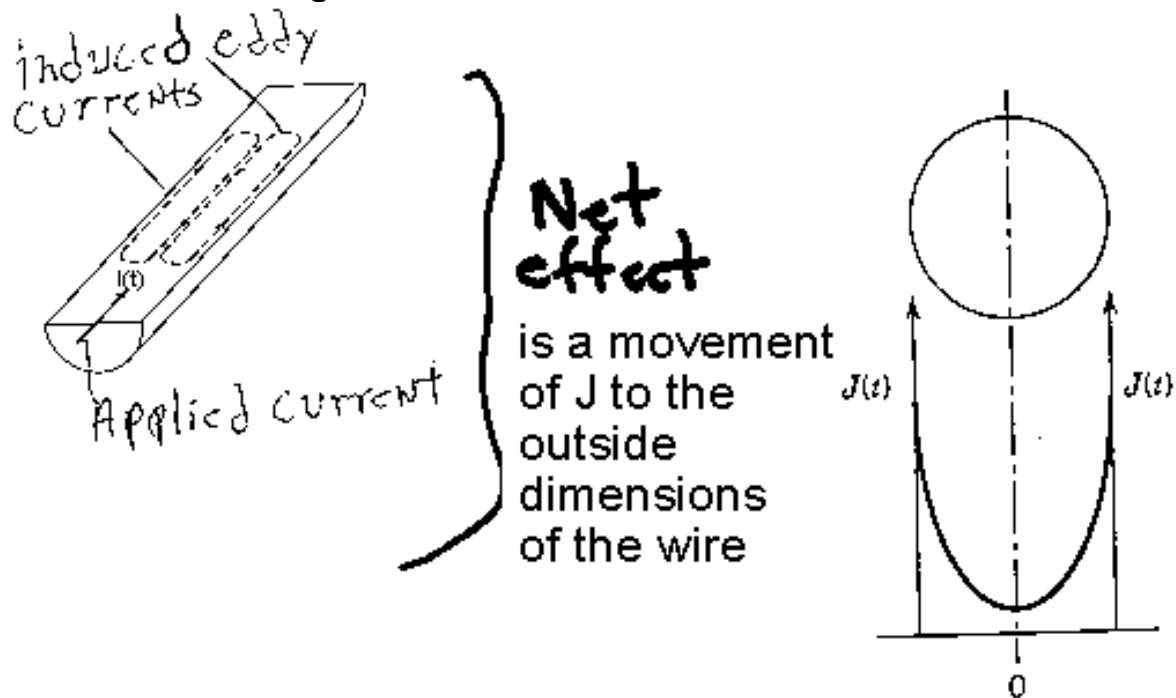


Skin Effect Model

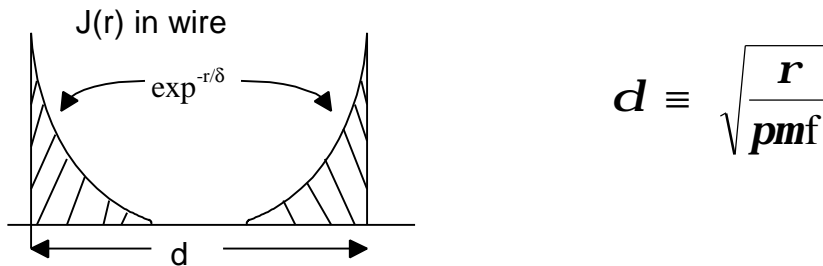
What happens when a high frequency ac current is put through? Does the effective resistance increase due to the large inductive impedances? Note that in the parallel combination of resistors the far right hand side R_i resistance's, which correspond to the center of the wire, are blocked by high inductive impedance at high frequencies. The far left R_i resistances, which correspond to the outer surface of the wire, are left unblocked by the inductive effects. The net effect on high frequency current distributions at center of the wire as f increases is to get zero net current. Thus in a wire of radius a , the current density profile $J(r)$ in units of (A/m^2) , will vary from dc, to a new profile at low f ac, to a different profile at high f ac. This is the skin effect for a single isolated wire only. Groups of wires wound in turns around cores will behave

differently as the H fields each wire sees will also vary.

In summary, at dc or low frequency ac, energy transfer back and forth to the wire inductance L_i , over time, is trivial compared to energy dissipated in the resistance. The current distributes itself uniformly through the wire from the surface to the center, to minimize the I^2R loss. But at high frequency, over the short time spans involved, I^2R loss is less than the energy transfer to and from L_i . Current flow then concentrates near the surface, even though the net resistance is much greater, in order to minimize inductive energy transfer to L_i . If we look at this strictly from a circuit point of view, at high frequency, the impedance of L_i near the surface blocks the current from flowing in the center of the wire. Hence, we get current distributions as shown below.



To a first approximation, J decays exponentially from the outer surface of the wire towards the center with a spatial decay constant δ , $J(r) \sim e^{-r/\delta}$ as shown. The diameter of the wire, d , as compared to the skin depth, δ , is the key parameter as this sets the spatial decay pattern of the current density, J , in the wire from the outer surface to the center of the wire. Calculation of δ is beyond this course but found in any electromagnetic fields text.

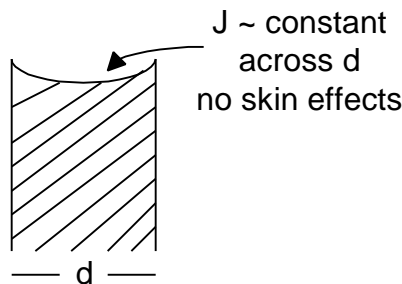


Using wire resistivity and wire permeability constants for Cu wire media we find a crude RULE OF THUMB for Cu wires:

$$d(\text{cu @100}^\circ\text{C}) = \frac{7.5 \text{ cm}}{\sqrt{f}}$$

Frequency	50	5 kHz	20 kHz	500 kHz
δ	10.6 mm	1.06 mm	1/2 mm	0.106 mm

At 500kHz wires bigger than 0.1 mm in diameter will show skin effects. In general, if $d(\text{wire}) \leq 2\delta$ (skin depth) we will not appreciably alter $J(r)$ in the small diameter wire by the skin effect as shown below. This is the case for AC transmission lines at 50-60 Hz.



Side comment on Skin Effects in modern Microelectronics Chips: Due to small wire size in microelectronics chips, typically less than one micron wide wires, the skin effect is not an issue. Consider possible skin effects in a new Intel μ Processor operating @ 1 GHz with copper interconnect wires. At this frequency the copper skin depth is

$$\delta(\text{GHz}) \sim 0.75 \text{ microns}$$

Thus only for those chip wires bigger than 1.5 microns in diameter skin effects will play a role. Note this means there is no real benefit going to thicker than 2μ copper metallization lines in the μ Processor wiring, as one might do especially for high AC currents in wires. DC power busses will still benefit from wider dimension wires. For copper interconnect, with wires around 1.5 microns we see a break point. Below this value no skin effect occurs. Above this value of wire line width BIG skin effects occur.

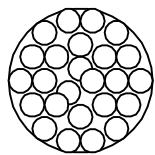
$d < 1.5 \mu$ wide signal bus $d > 1.5 \mu$ wide wire

$$R_{dc} = R_{ac}$$

$$R_{ac} \gg R_{dc}$$

This is counter intuitive for some people who expect less skin effect problems for thicker wires. In reality it is just the opposite.

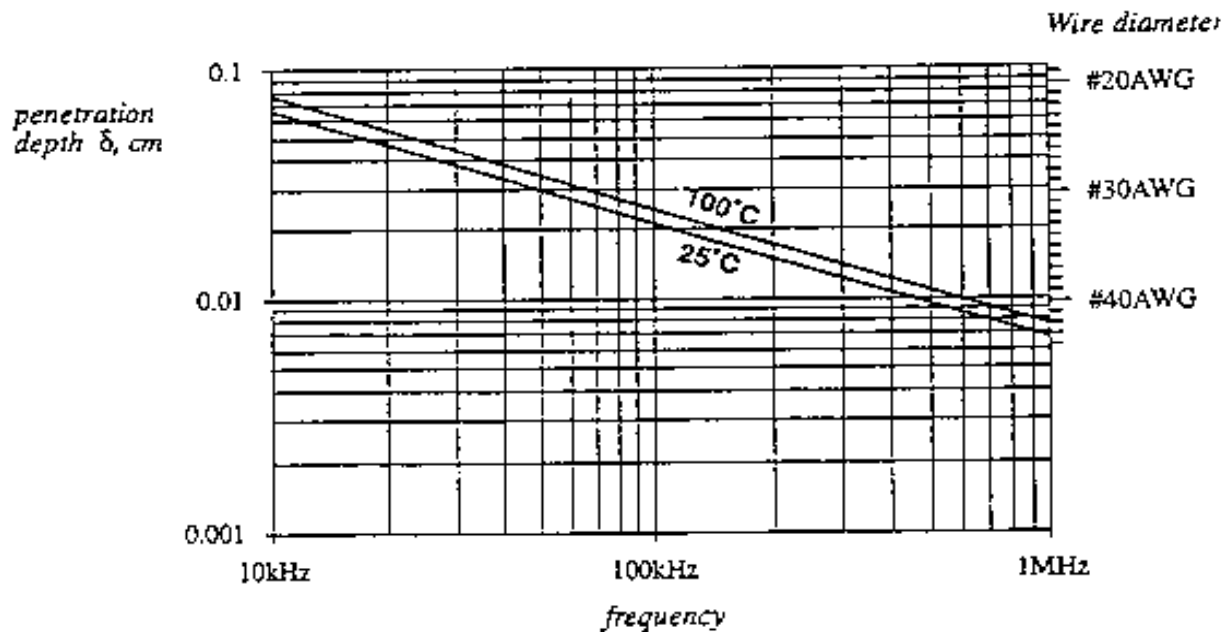
As a consequence a bundle of thin wires is the way to make an effective large diameter wire in macroelectronics as practiced in power electronics. This low AC loss type of wire is commercially available and is termed LITZ wire



← Litz wire

345 strands of #35 wire
in parallel strands are woven together.
Each strand of Litz has a
diameter ~ 200 microns

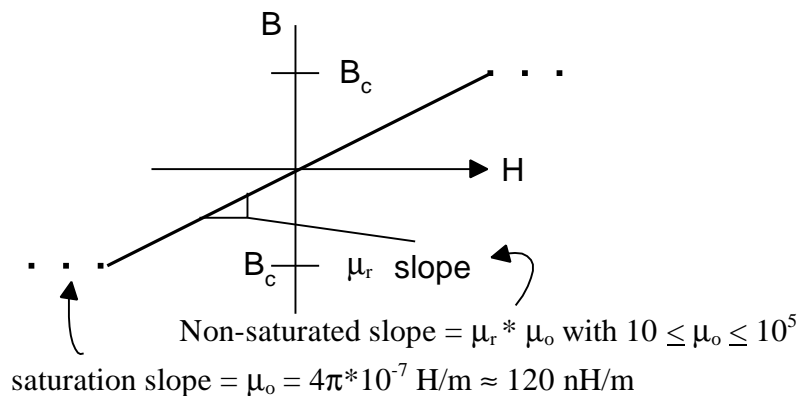
Copper wires come in specific wire sizes. For Cu wires the δ vs f plots can be compared to the standard AWG # and associated wire diameter to better see when $d > 2\delta$ occurs and skin effects are LARGE. With these wires we interconnect the converter circuits containing solid state devices and magnetic devices. An informative plot of skin depth versus frequency benefits greatly with horizontal notations of where the various wire sizes lie. We don't want to specify a larger diameter wire to handle large currents only to find out a smaller diameter wire would do the job just as well. Page 12 plots of δ versus f with wire diameter as a horizontal bar, when δ hits wire diameter, helps one keep this relative size relationship in mind.



C. Magnetic Materials B-H Curves and Losses

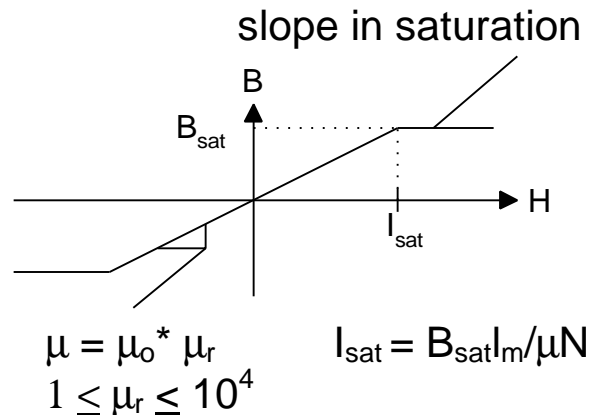
1. Basics

The role of an ideal magnetic core is to transfer energy from one wire winding to another wire winding in a loss free way. First a brief review of B-H curves that includes saturation effects.

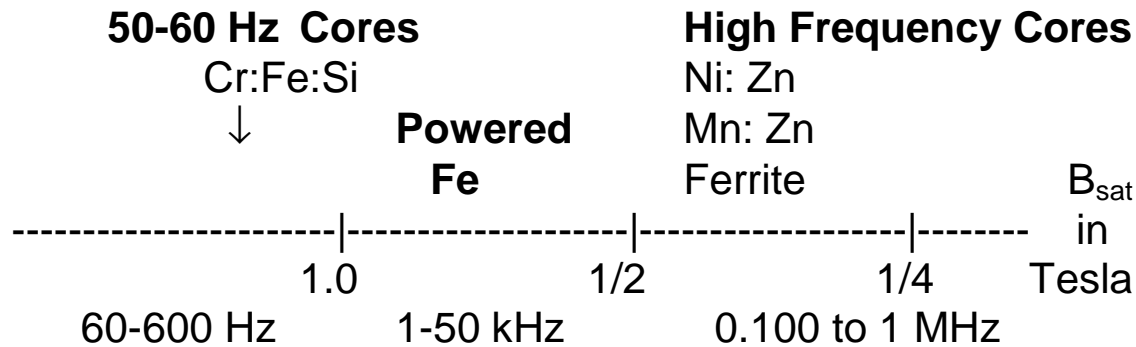


From Faraday's Law $V = Nd\phi/dt$ which says that $\phi = \int Vdt/N$. Hence the B or y -axis corresponds to $\int Vdt$, the integral of the applied voltage to the wire coil. H or the x -axis also represents current in the wire coil, so that electrically B-H corresponds to $\int Vdt$

vs I. For a series of wire turns wound around a wooden or paper form, no B-H saturation occurs and we only see the μ_0 slope at all values of the B-H curve. Using the fact $(1/L)\int Vdt = i$, a **small B-H slope corresponds to a small value inductance**. If we wish to make a bigger L in the same volume or the same L in a smaller volume then we employ a core magnetic material with $\mu_r \gg 1$. By winding turns around a high μ_r core the B-H slope rises by a factor μ_r and so does the inductance of wires wound around that core to applied voltages, V. In short that's why big inductors in a small volume require a magnetic media or core to be employed. The price we may have to pay is non-linear saturation effects. That is big μ_r occurs only below I(critical) or H(critical). At H(critical) we have reached B_{sat} of the chosen core and the L value decreases.



Below we plot typical values of B_{sat} for various cores as well as their low loss(~1 %) frequency operation range.



Note that while B_{sat} varies with material choice for the core,

unfortunately, B_{sat} for various cores **only** varies over ONE ORDER OF MAGNITUDE. The science of high permeability cores needs further work to catch up to the needs of technology.

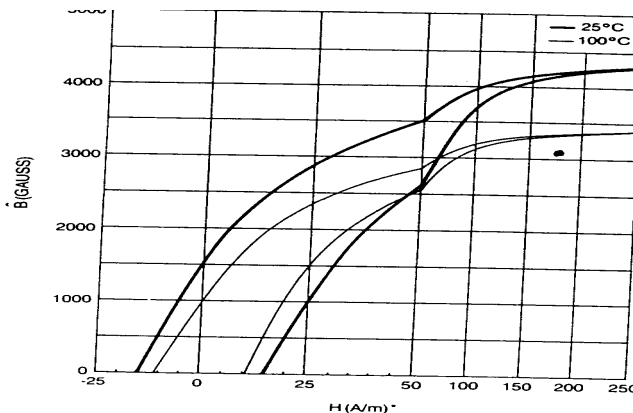
Common Materials choices and Manufacturers are listed below.

Core Materials		
Manufacturer	Core Material (Ferrites)	
	<100 kHz	<1 MHz
Magnetics, Inc.	F, T, P	F, K, N
TDK	P ₇ C ₄	P ₇ C ₄₀
Ferroxcube	3C8	3C85
Siemens	N27	N67

Next we list practical guides to the actual B_{SAT} limits in cores versus the frequency of the current in the coils wound around the core. Note that in practice the full B-H curve cannot be employed at higher frequencies (above 1MHz) due to excessive losses.

Flux Density Limits vs. Frequency	
Frequency	Maximum Operational Flux Density (B_{max})
<50 kHz	$0.5B_{sat}$
<100 kHz	$0.4B_{sat}$
<500 kHz	$0.25B_{sat}$
<1 MHz	$0.1B_{sat}$

Elevated core temperatures due to heating effects will lower B_{SAT}

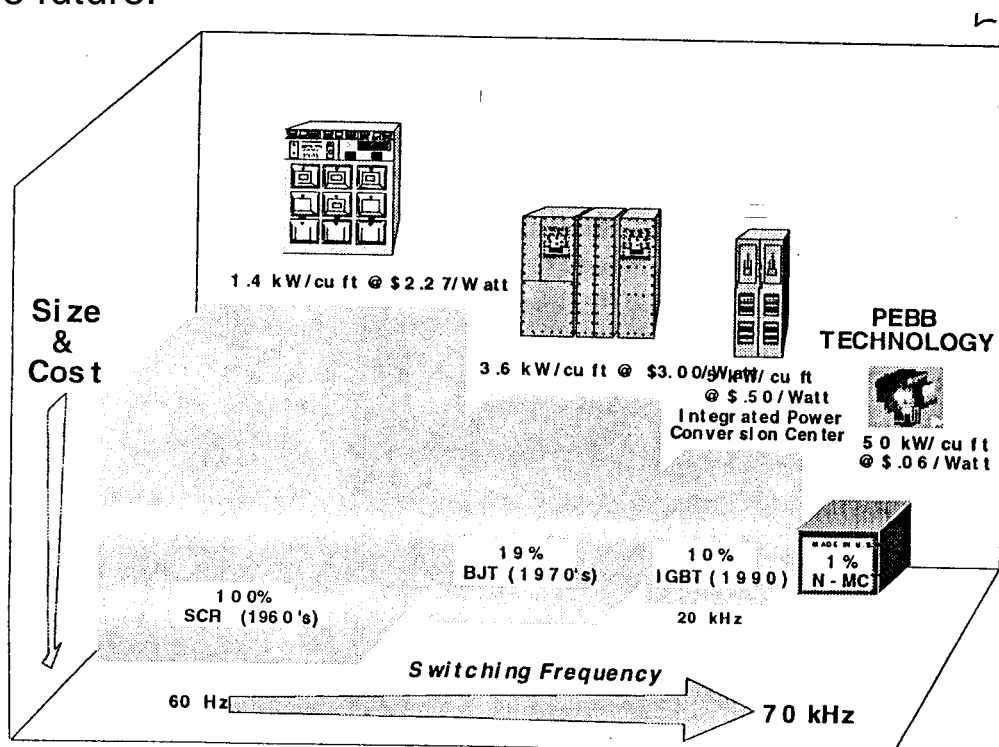


* 100 A/m is 1.25 oersted

degradation of B_{sat} with core temperature (3C8 material shown). (Courtesy of Philips Components.)

This means for the same size applied voltage to one set of wire coils we could employ 10^5 less flux ϕ in the core by operating at 10^5 higher frequency. In short, we can shrink the area of the core (size / weight) by this same factor and still reach V_{out} on a second set of coils wound on the same core. This allows spatially small size transformers to control large powers.

The above change in the operating frequency of magnetic materials was crucial to the shrink in the size of magnetic components over the past 20 years and in conjunction with the advances in solid state switches made kW supplies go from four drawer file cabinet size to shoe box size. Clearly, the advancements in magnetics, as well as those in solid state switches, both pace and limit the progress in switched mode power supplies that we have seen over the past decade and into the future.



2. Losses in Magnetic Cores

a. Overview

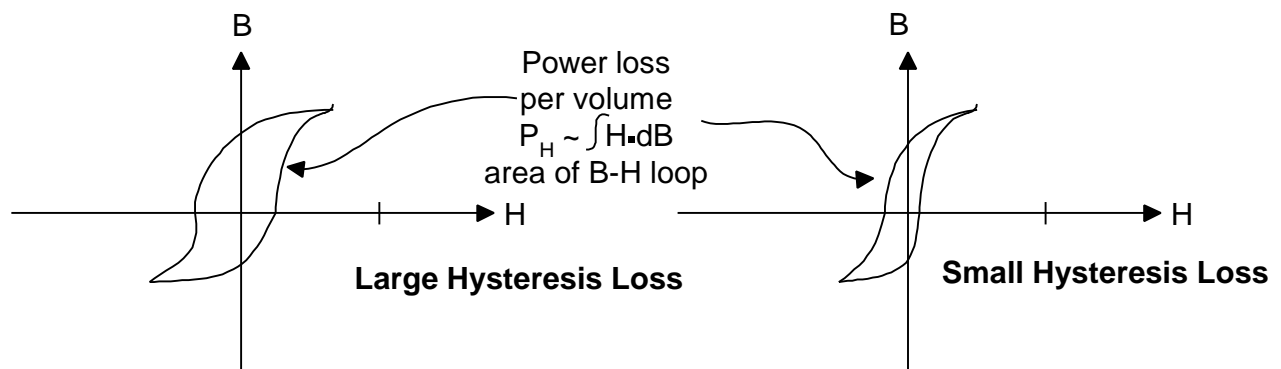
Core loss is the second most important core limitation in most PWM converter applications after active switch losses. . Energy losses cause cores to heat up during operation. Losses arise from both hysteresis effects and eddy currents.

In metal alloy cores, eddy current loss dominates above a few hundred Hertz but this is not the case for ferrites until much higher frequency. For acceptable ferrite core losses, flux density swing ΔB must be restricted to much less than B_{sat} , which prevents the core from being utilized to its full capability. At low frequencies, ferrite core loss is almost entirely hysteresis loss. For today's power ferrites operating at high frequency, eddy current loss overtakes hysteresis loss at 200-300kHz..

Core manufacturers usually provide curves showing core loss as a function of flux swing and frequency, combining hysteresis and eddy current losses. Core loss is usually expressed in mW/cm^3 , sometimes in kW/m^3 (actually equal : $1\text{mW}/\text{cm}^3 = 1\text{KW}/\text{m}^3$). Now lets discuss quantitatively two core losses: hysteresis and eddy currents.

a. Hysteresis Loss

Here the **fatness or thinness of the B-H curve** when excited by ac currents at frequency f is key. This loss also depends on $B(\text{peak})$ not $B(\text{average})$ or B_{RMS} . Usually Loss varies as $B(\text{peak})^m$ where $m = 2-3$.

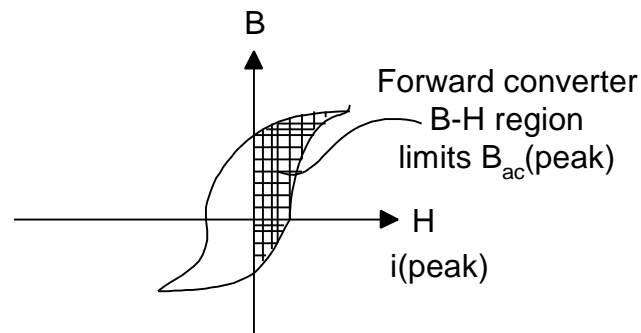


Notice also this loss occurs over each ac cycle so total loss is proportional to applied frequency f .

Total power Loss $\equiv \int H \cdot B$ * f * Volume of Core
 energy loss
 per volume
 per cycle

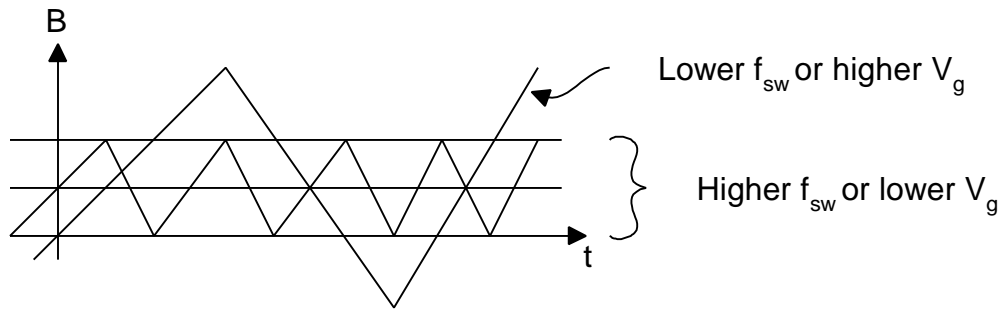
$$P_{\text{loss}}(\text{Hysteresis}) \sim \left[\begin{array}{c} \text{core material type} \\ \text{and} \\ B_{\text{ac}}(\text{peak}) \end{array} \right] * f * \text{core size}$$

$B^m(\text{peak})$ is less if core area is bigger, yet the total loss decreases as the core size gets smaller. So optimum geometry's for cores that minimize total loss do exist. These optimums will also have to include core cooling effects. Circuit topology also effects the B-H curve. For example how many quadrants of B-H are employed depends on whether the wire current that is wrapped around a core is unipolar or bipolar driven by the converter switches. "L" in a forward converter, for example, is only in quadrant 1 of the B-H because the diode allows only uni-directional current. Hence, we expect higher B_{sat} and losses may be bigger than for bipolar excitation where B_{SAT} is lower.



Clearly in a forward converter both the f_{sw} and V_g choices also

effect the total loss via the $\frac{V_g}{L}$ (A/sec) ramp time as set by choice of T_{SW} . This in turn sets the maximum current which effects $\int E dt \sim B(\text{peak})$ as shown below. We do not wish to exceed B_{SAT} at any time.



Note also it depends if there is a net dc current level that causes B to vary about a quiescent B_{DC}

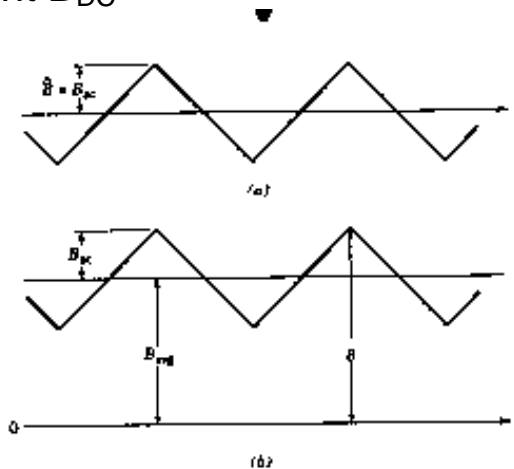
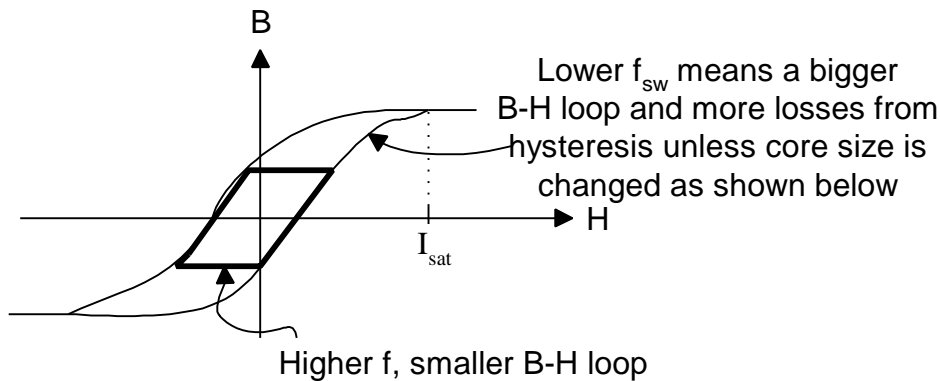


Figure 30-1 Magnetic flux density waveforms having (a) no time average and (b) with a time average.



$I_{\text{sat}} \equiv \frac{B_{\text{sat}} l_m}{mN}$ Where N is the # of copper wire turns around the

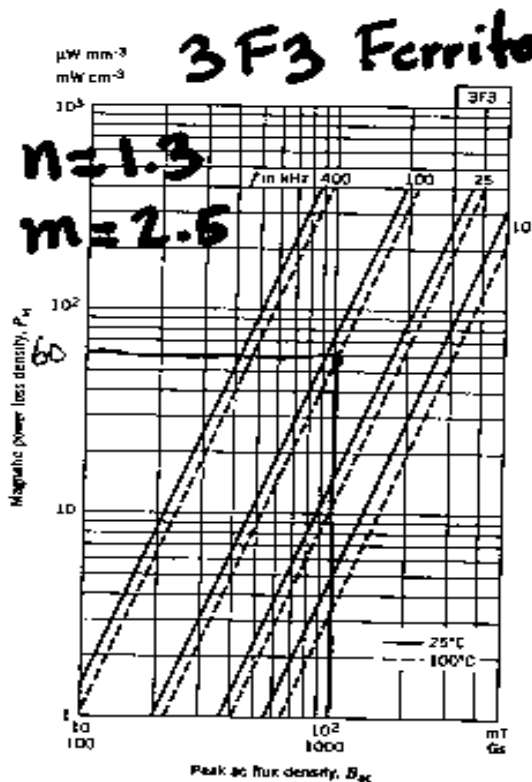
core and l_m is the perimeter length of the chosen core.

Lower V_g or Higher f means Smaller $B_{\text{ac}} \leftarrow \rightarrow$ Smaller Core Size required for same total loss.

In summary, we find: $P(\text{hysteresis}) = K f (B_{\text{ac peak}})^m$

Below we show data of **loss versus B(peak)** on a log-log scale so that an m value is a straight line. Changing frequency changes the loss as shown. In this case loss varies as $f^{1.3}$. $P(\text{hysteresis})$ loss is also limited by thermal heating which changes m to higher values at higher temperatures. A rule of thumb is $T_{\text{max}} 100 - 200^\circ\text{C}$ for the ferrite. This upper temperature is or similar to T_{max} for safe semiconductor switch operation or for the wire insulation to remain undamaged.

$P(\text{hysteresis max})$ is typically $\approx 200 - 500 \frac{\text{mW}}{\text{cm}^3}$

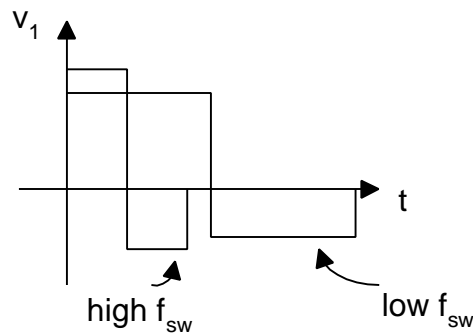


Consider changing the operating f from 100 to 400 kHz on the wire current wound around 3F3 ferrite for use as an inductor core. Start with @100 kHz and $B_{ac} = 100$ mT the power loss density in the core is, $P = 60$ mW/cm³. By increasing operating f to 400 KHz the power loss density is 10* greater, $P = \frac{600\text{mW}}{\text{cm}^3}$, and not simply 4 times greater due to non-linear core heating.

We repeat the warning that in Core Loss vs. Flux Density curves, the horizontal axis labeled “Flux Density” usually represents peak flux density. In applications, peak-to-peak flux swing, ΔB , is calculated from Faraday’s Law, where $\int E dt =$ applied Volt-seconds, $N =$ turns, and $A_c =$ core cross-sectional area:

$$\Delta B = (1/NA_c) \int V_1 dt$$

The total flux swing, ΔB , is sometimes twice the peak flux swing referred to in the core loss curves as “Flux Density” if we choose unipolar versus bipolar core excitation. Therefore, use care to employ either $\Delta B/2$ or ΔB to enter into the core loss curves.

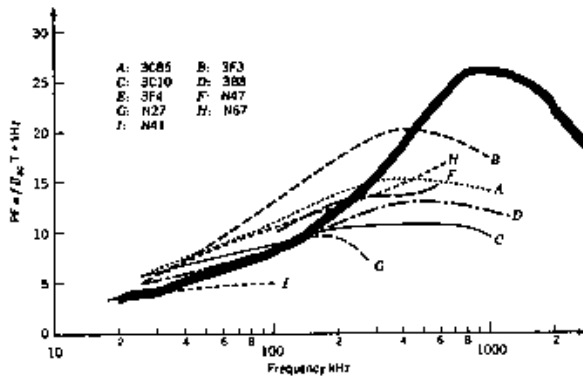


Aside: Later we will show that for transformers (but not for inductors)

$$\text{V-A rating transformer} \quad \longleftrightarrow \quad f * B_{ac} \text{ product}$$

The $f \times B_{ac}$ **product** versus applied f peaks at a characteristic frequency unique to each magnetic material. 3F4 Material is considered one of the best for high f operation as shown in the

plot below where the product $f \times B$ peaks near 1MHz. One is reminded of the gain bandwidth product for electronic amplifiers.



This plot shows the fact that E(3F4 material) has 2.5 * more V-I capability than 3C10.

Figure 30-3 Empirical performance factor $PF = fB_{dc} T + 1/4f$ versus frequency for various ferrite core materials. Measurements are made at a power density $P_{core} = 100 \text{ mW/cm}^2$.

All present ferrites have low $f \times B$ products above 2 MHz. MHz is today's $f_{sw}(\text{max})$ primarily due to magnetic materials limitations. This is an area ripe for a scientific breakthrough.

b. Eddy Current Effects in Various Core Materials

The eddy current core loss has entirely unique physical origins and acts in parallel with that associated with hysteresis loss.

1. Overview

By Faraday's law we induce a voltage loop in a core material with $v \sim N \omega \phi$. This loop voltage induces a current via $V/R(\text{core})$.

The electrical resistivity of the core material can vary:

Traditional Fe:Si

low ρ

easy to induce bulk currents in the core

Which creates Mainly eddy current loss = $I^2 R$ \uparrow

Ceramic Ferrite

high ρ

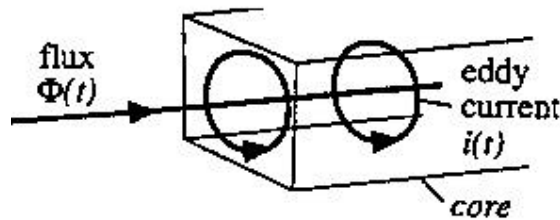
difficult to induce eddy currents

Only hysteresis loss occurs

Most cores have both hysteresis and eddy current losses. The resistivity of the core is crucial because magnetic fields that vary in time create an induced voltage in the core as shown below.

This current flowing in the core material causes losses as either V^2/R or I^2R .

b. The induced voltage in the core bulk by Faraday's Law is $V \sim Nw\dot{\phi}$ for harmonic varying flux perpendicular to the core area. The voltage is a circular potential causing circular currents to flow in the core as shown below. Also loss goes as f^2 for eddy currents.



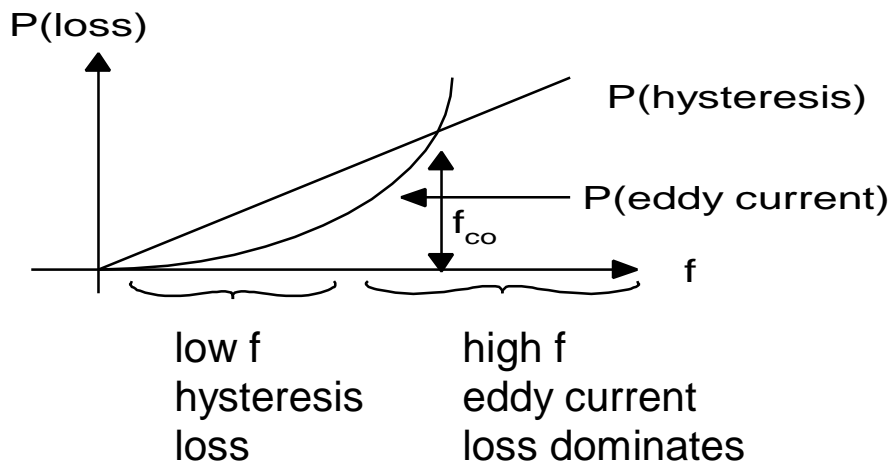
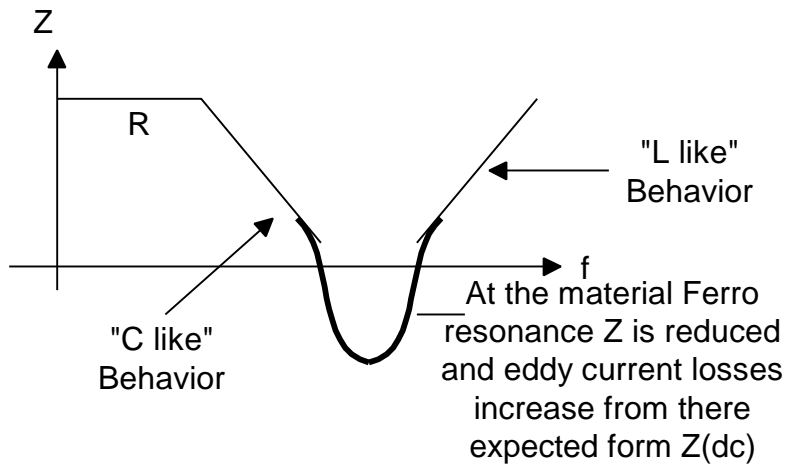
$$P(\text{eddy current}) \approx i^2R$$

$$i_{\text{eddy}} \approx \frac{V}{R} \sim \frac{fB}{R}$$

$$P(\text{eddy loss}) \sim \frac{f^2 B^2}{R} \Rightarrow \text{Fe:Si has big}$$

loss since $R(\text{core})$ is small. Ferrite has low eddy current loss since $R(\text{ferrite})$ is large. Careful of ferrite resonance's

If you put ferrite core on vector voltmeter to measure $Z(\text{ferrite})$ you will find a $Z(\text{min})$ at $f(\text{resonance})$. In general, for ferrite ρ is large, $1000 \Omega\text{-cm}$, So eddy current loss for a ferrite core is low except at ferro-resonance. One must know the resonant frequency for the ferrite core material you choose to work with. Again, low impedance means V^2/R has a maximum. At any rate at some applied frequency the eddy current loss, which varies as f^2 will exceed the hysteresis loss which varies as f . This is shown on page 24.



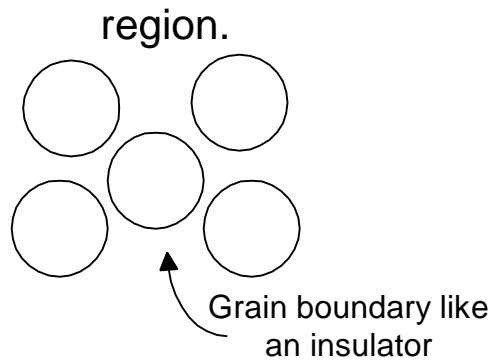
f_{co} , $f(\text{crossover})$ is typically 100 kHz.

At low f
hysteresis dominates

We can reduce eddy current by laminating the core to block eddy currents unless you use high ρ ferrite cores. At very high $f(\text{MHz})$
 $P_{\text{loss}}(\text{ferrites}) \sim f^3$

High Frequency Cores

No high ρ ferrites exist with simultaneous high B_{sat} in the MHz



TDK H7C4 (high f cores) have very small grain size (1micron)

$$V(\text{eddy}) \sim f B$$

$$i(\text{eddy}) \sim \text{At high frequency the}$$

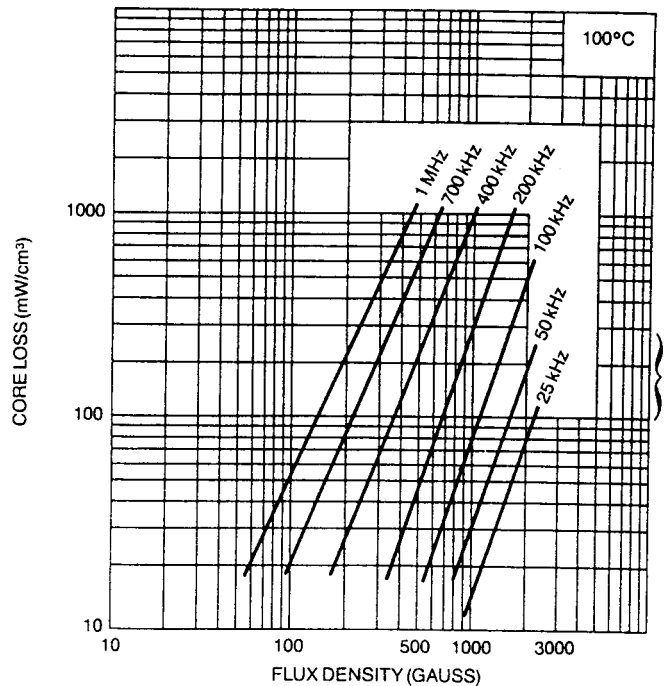
If capacitance between grain boundaries cause displacement current flow not just bulk resistivity, R . Then in a high frequency core to a first approximation if the "C" impedance between grains dominates

$$i_{\text{eddy}} \sim \frac{V(\text{eddy})}{Z} = \frac{V(\text{induced})}{\frac{1}{\omega C} \parallel R} = \omega C V \sim \omega^2 B C$$

With $V = \omega B$.

$$P_{\text{eddy}} \sim V(\text{eddy}) i(\text{eddy}) \sim \omega^3$$

In summary, the total core loss from all magnetic effects will vary with both the flux density, B and the applied frequency of the currents in the wires wound around the core as follows:



* Recommended operating range for loss limited designs.

Flux density for bipolar excitation is B_p .

Flux density for unipolar excitation is $\frac{B_{p-p}}{2}$.

Curves showing volumetric core loss vs. frequency and B_{max} (3C8 material shown). (Courtesy of Philips Components.)

The small region in the power loss-B curve with a star, *, represents the suggested operation region to achieve the condition that the core losses are well below the 1-5% of the transmitted power through the core.

What is left for later lectures is what the lost energy does to the temperature of the core. Later we will balance energy loss input to the core with energy lost to the environment by various cooling paths and end up with the equilibrium temperature of the core under operation conditions. This temperature should never exceed 100 degrees Centigrade to preserve both the core and the wire insulation.

FOR HWS# 4 and 5 do the following:
 HW#4 Erickson Chapter 12 problems 6 and 8 and in class questions as well as lecture questions

Hw# 5 Erickson Chapter 13 problem 3 and in class questions as well as lecture questions