Transformer & Magnetic Design

- Transformer Core Materials & Geometries
- Core Losses
- Peak Flux Density Selection
- Maximum Transformer Core Output Power
- Flyback Transformer (Magnetics)
- Powdered Molypermalloy (MPP) Cores
- Transformer Copper Losses
Transformer & Magnetic Design

Why do we need a transformer?

- Isolation for off-line operation
- Obtain the output voltage for a given duty cycle
- Multiple outputs
Selection of Topologies

Consideration of

- the voltage stress of the power switch at high line
- the current stress of power switch at maximum output
- the operating frequency and transformer core size
Transformer Core Materials

Ferrites
- Mixture of iron, manganese or zinc oxides
- Ceramic ferromagnetic material (brittle!)
- High resistance, no eddy current loss
- Core loss is mainly due to hysteresis loss
Transformer Core Geometries

Pot cores (up to 125W)
- Almost entirely enclosed by ferrite material therefore excellent EMI performance
- Narrow notch for coil leads
- Restrict to low voltage and current applications
Transformer Core Geometries

EE cores (5W to 10kW)
- Most widely used core
- No narrow notch restricting coil leads
- Poor EMI performance
- Good thermal performance due to unimpeded airflow
Transformer Core Geometries

EC & ETD cores
- Improved EE core
- Round centre leg
- Reduce the mean length of a turn for a given core area by 11%
- Lower Copper loss
Transformer Core Geometries

RM or “Square” cores

- Compromise between a pot and an EE core
- Good EMI performance
- No narrow notch
- Unimpeded airflow
Transformer Core Geometries

PQ cores
- Optimum ratio of volume to radiating surface and coil winding area
- Minimizing temperature rise for a given power
- Minimizing volume for a given power
Transformer Core Geometries

LP cores
- Designed for low profile applications
- Minimizing leakage inductance with long centre legs
Transformer Core Geometries

UU and UI cores

- Mainly for high-voltage or ultra-high power applications
- Large window area permits thick wire size
- Large leakage inductance
MKS Units

Flux $\Phi$ in weber = $B \times \text{Area in } m^2$

Flux Density $B$ in tesla

Magnetic Field Strength $H$ in A/m

$H = NI/l$

where magnetic path length $l$ is in m

- And $B = \mu_0 H$ in free space

where $\mu_0$ is $4\pi \times 10^{-7} \text{ H/m}$
CGS Units

Flux $\Phi$ in maxwell = $B \times \text{Area in cm}^2$
Flux Density $B$ in gauss
Magnetic Field Strength $H$ in oersted
$H = \frac{(0.4 \pi NI)}{l}$

where magnetic length $l$ is in cm

Why CGS? Because $B = H$ numerically in free space
Conversion Between Units

- 1 weber = $10^8$ maxwell
- 1 tesla = $10^4$ gauss
- 1 A/m = 0.0126 oersted

Faraday’s Law of induction

$V = N \left( \frac{d\Phi}{dt} \right)$ in MKS

$V = N \left( \frac{d\Phi}{dt} \right) \times 10^{-8}$ in CGS, i.e., if $\Phi$ is calculated from gauss $\times$ cm$^2$
At 100°C, complete saturation occurs at a flux density of 3200 G. \(1\text{mT} = 10\text{G}\)

At 100°C, residual flux density is about 900 G. 3F3
Core Losses

Mainly due to hysteresis loss; expressed in mW/cm³

![Graph showing peak flux density for different frequencies and flux densities.]

- **Peak flux density**
  - 1600G
  - 1200G
  - 800G

- **Frequencies**:
  - 20kHz
  - 50kHz
  - 100kHz
  - 200kHz

- **Note**: The graph illustrates the peak flux density at different frequencies and flux densities, indicating the core losses in the material.
Core Losses

Core loss $\propto (B_{\text{max}})^{2.7}$
Core loss $\propto f^{1.7}$

- Value of core losses are usually given for bipolar magnetic circuits (first and third-quadrant operation) by various manufacturers
- For unipolar circuits, e.g., forward, divide by 2
# Table of Core Losses

Core Losses for Various Core Materials at Various Frequency and Peak Flux Densities at 100°C.

<table>
<thead>
<tr>
<th>Frequency, kHz</th>
<th>Materials</th>
<th>Core loss, mW/cm³ for various peak flux densities, G</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>1600</td>
</tr>
<tr>
<td>20</td>
<td>Ferroxcube 3C8</td>
<td>85</td>
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<td></td>
<td>Ferroxcube 3C85</td>
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<td></td>
<td>Ferroxcube 3F3</td>
<td>28</td>
</tr>
<tr>
<td></td>
<td>Magnetics Inc.-R</td>
<td>20</td>
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<td></td>
<td>Magnetics Inc.-P</td>
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<tr>
<td></td>
<td>TDK-H7C1</td>
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<td></td>
<td>TDK-H7C4</td>
<td>45</td>
</tr>
<tr>
<td></td>
<td>Siemens N27</td>
<td>50</td>
</tr>
</tbody>
</table>
Peak Flux Density Selection

Although BH loop is still linear up to 2000G, peak flux density should be chosen at 1600G such that power switch will not be destroyed due to saturation which may occur with fast load change if peak flux density is too high.
Peak Flux Density Selection

- How to detect the occurrence of saturation in practice?

Current Waveform

Saturation starts
Maxi. Transformer Core Output Power

Forward Topology

\[ P_o = \frac{(0.5 B_{\text{max}} f A_e A_b)}{D_{\text{cma}}} \text{ Watts} \]

- Peak flux density \( B_{\text{max}} \) in Gauss \( \quad \text{1mT} = 10 \text{G} \)
- Switching frequency \( f \) in kHz \( \quad \text{1A/m} = 0.0126 \text{ Oersted} \)
- Core area \( A_e \) in \( \text{cm}^2 \)
- Bobbin winding area \( A_b \) in \( \text{cm}^2 \)
- \( D_{\text{cma}} \) in circular mils per rms ampere (Inverse of current density)
What is \( D_{\text{cma}} \) in circular mils per rms ampere?

- Area per unit ampere
- \( 500 \text{ cma} \) means that a wire with a diameter of \( \sqrt{500} \text{ mils} \) (1 mil = 1/1000 of an inch) carries 1A rms

\[
\sqrt{500} = 22.3 \text{ mil} = 0.566 \text{ mm}
\]
Maxi. Transformer Core Output Power

How did we come up with this equation?

\[ P_o = \left(0.5 \times B_{\text{max}} \times f \times A_e \times A_b\right) / D_{\text{cma}} \quad \text{Watts} \]

Voltage \quad Current

Assume that

- Efficiency is 0.8
- Space factor SF is 0.4
- Duty cycle is 0.4
Maxi. Transformer Core Output Power

What is space factor?

- The fraction of bobbin winding area occupied by the winding of primary and secondaries

Why SF is 0.4?
Maxi. Transformer Core Output Power

We can increase the output power by

- Higher Efficiency (Mission Impossible!)
- Higher SF (Difficult!)
- Higher duty cycle (Yes, using different reset arrangement, but higher voltage stress on power switch)
- Higher switching frequency (Yes, but higher switching and core losses)
Primary & Secondary Windings

Forward Topology

\[ N_p = \frac{40000 \, V_{dc}}{(A_e \, \Delta B \, f)} \text{ Turns} \]

- Core area \( A_e \) in \( \text{cm}^2 \)
- \( \Delta B = B_{\text{max}} \) (Use 1600 Gauss)
- Switching frequency in kHz
- DC input voltage \( V_{dc} \) in volts

Assume that duty cycle is 0.4
Primary & Secondary Windings

Forward Topology

\[ V_o \approx 0.4 \ V_{dc} \left( \frac{N_s}{N_p} \right) \]

\( N_s \) can be easily found

- Wire size of primary winding is \((988 \ P_o)/V_{dc}\) in circular mils

- Wire size of secondary winding is \(316 \ I_o\) in circular mils

Assume that duty cycle is 0.4 and efficiency is 0.8
Maxi. Transformer Core Output Power

**Push-Pull Topology**

\[ P_o = \frac{(B_{\text{max}} f A_e A_b)}{D_{\text{cma}}} \text{ Watts} \]

- Peak flux density \( B_{\text{max}} \) in Gauss
- Switching frequency \( f \) in kHz
- Core area \( A_e \) in cm\(^2\)
- Bobbin winding area \( A_b \) in cm\(^2\)
- \( D_{\text{cma}} \) in circular mils per rms ampere (Inverse of current density)
Maxi. Transformer Core Output Power

**Push-Pull Topology**

\[ P_o = \frac{(B_{max} f A_e A_b)}{D_{cma}} \text{ Watts} \]

- Output is two times compared to that of Forward Topology

Why?

- Because the flux change is doubled
- Core is warmer due to double in core losses
Primary & Secondary Windings

**Push-Pull Topology**

\[ N_p = \frac{(40000 \ V_{dc})}{(A_e \ \Delta B \ f)} \] \text{ Turns}

- Core area \( A_e \) in cm\(^2\)
- \( \Delta B = 2 \ B_{\text{max}} \) (Use 2 \times 1600 Gauss)
- Switching frequency in kHz
- DC input voltage \( V_{dc} \) in volts

Assume that duty cycle is 0.4
Primary & Secondary Windings

Push-Pull Topology

\[ V_o \approx 0.8 \ V_{dc} \ (N_s/N_p) \]

\( N_s \) can be easily found

- Wire size of primary winding is \( \frac{494 \ P_o}{V_{dc}} \) in circular mils
- Wire size of secondary winding is \( 316 \ I_o \) in circular mils

Assume that duty cycle is 0.4 and efficiency is 0.8
Maxi. Transformer Core Output Power

**Half- or Full-Bridge Topologies**

\[ P_o = \left(1.4 \ B_{\text{max}} \ f \ A_e A_b \right)/D_{\text{cma}} \text{ Watts} \]

- Peak flux density \( B_{\text{max}} \) in Gauss
- Switching frequency \( f \) in kHz
- Core area \( A_e \) in \( \text{cm}^2 \)
- Bobbin winding area \( A_b \) in \( \text{cm}^2 \)
- \( D_{\text{cma}} \) in circular mils per rms ampere (Inverse of current density)

\( 1 \text{mT} = 10 \text{G} \)
Maxi. Transformer Core Output Power

Half- or Full-Bridge Topologies

- I thought a Full-Bridge delivers twice the output power of a Half-Bridge!! (NOT TRUE in transformer)

- If the same core size is used both for Half-Bridge and Full-Bridge, no. of primary turns must be doubled in Full-Bridge which leads to a reduction in wire size, current, by half. (Equal power)
Primary & Secondary Windings

Half-Bridge Topology

\[ N_p = \frac{(20000 \ V_{dc})}{(A_e \ \Delta B \ f)} \text{ Turns} \]

- Core area \( A_e \) in cm\(^2\)
- \( \Delta B = 2 \ B_{\text{max}} \) (Use 2 \times 1600 Gauss)
- Switching frequency in kHz
- DC input voltage \( V_{dc} \) in volts

Assume that duty cycle is 0.4
Primary & Secondary Windings

Half-Bridge Topology

\[ V_0 \approx 0.4 \, V_{dc} \left( \frac{N_s}{N_p} \right) \]

- \( N_s \) can be easily found
- Wire size of primary winding is \( \frac{1397 \, P_o}{V_{dc}} \) in circular mils
- Wire size of secondary winding is \( 316 \, I_o \) in circular mils

Assume that duty cycle is 0.4 and efficiency is 0.8
Primary & Secondary Windings

Full-Bridge Topology

\[ N_p = \frac{(40000 \ V_{dc})}{(A_e \ \Delta B \ f)} \text{ Turns} \]

- Core area \( A_e \) in cm\(^2\)
- \( \Delta B = 2 \ B_{\text{max}} \) (Use 2 \( \times \)1600 Gauss)
- Switching frequency in kHz
- DC input voltage \( V_{dc} \) in volts

Assume that duty cycle is 0.4
Primary & Secondary Windings

Full-Bridge Topology

\[ V_o \approx 0.8 \ V_{dc} \ (N_s/N_p) \]

\( N_s \) can be easily found

- Wire size of primary winding is \((699 \ P_o)/V_{dc}\) in circular mils
- Wire size of secondary winding is \(316 \ I_o\) in circular mils

Assume that duty cycle is 0.4 and efficiency is 0.8

Two current pulses per period

\[ 500 \times \frac{1}{0.8} \times \frac{1}{0.8} \times \sqrt{0.8} \]
# Table of Maxi. Output Power

Maximum Available Output Power in Forward Converter Topology

<table>
<thead>
<tr>
<th>Core</th>
<th>$A_e$, cm²</th>
<th>$A_b$, cm²</th>
<th>$A_eA_b$, cm⁴</th>
<th>20kHz</th>
<th>30 kHz</th>
<th>50 kHz</th>
<th>80 kHz</th>
<th>100 kHz</th>
<th>150 kHz</th>
<th>200 kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>814E250</td>
<td>0.202</td>
<td>0.171</td>
<td>0.036</td>
<td>1.1</td>
<td>1.7</td>
<td>2.8</td>
<td>4.4</td>
<td>5.5</td>
<td>8.3</td>
<td>11.1</td>
</tr>
<tr>
<td>813E187</td>
<td>0.225</td>
<td>0.329</td>
<td>0.074</td>
<td>2.4</td>
<td>3.6</td>
<td>5.9</td>
<td>9.5</td>
<td>11.8</td>
<td>17.8</td>
<td>23.7</td>
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<tr>
<td>813E343</td>
<td>0.412</td>
<td>0.359</td>
<td>0.148</td>
<td>4.7</td>
<td>7.1</td>
<td>11.8</td>
<td>18.9</td>
<td>23.7</td>
<td>35.5</td>
<td>47.3</td>
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<tr>
<td>812E250</td>
<td>0.395</td>
<td>0.581</td>
<td>0.229</td>
<td>7.3</td>
<td>11.0</td>
<td>18.4</td>
<td>29.4</td>
<td>36.7</td>
<td>55.1</td>
<td>73.4</td>
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<tr>
<td>782E272</td>
<td>0.557</td>
<td>0.968</td>
<td>0.639</td>
<td>17.3</td>
<td>25.9</td>
<td>43.1</td>
<td>69.0</td>
<td>86.3</td>
<td>129.4</td>
<td>172.5</td>
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<tr>
<td>E375</td>
<td>0.810</td>
<td>1.149</td>
<td>0.931</td>
<td>29.8</td>
<td>44.7</td>
<td>74.5</td>
<td>119.1</td>
<td>148.9</td>
<td>223.4</td>
<td>297.8</td>
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<tr>
<td>E21</td>
<td>1.490</td>
<td>1.213</td>
<td>1.807</td>
<td>57.8</td>
<td>86.8</td>
<td>144.6</td>
<td>231.3</td>
<td>289.2</td>
<td>433.8</td>
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<td>1.781</td>
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<td>783E776</td>
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<tr>
<td>E625</td>
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<td>3.206</td>
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<td>153.9</td>
<td>256.5</td>
<td>410.3</td>
<td>1025.9</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Transformers (Forward-Type Converters)

- In Forward-type converters, when the power switch is ON, primary current flows into a dot and secondary current flows out of a dot.
- Primary and secondary mmfs in transformer are balanced.
Transformers (Forward-Type Converters)

- Mmfs are balanced except magnetization inductance in the primary winding of a Forward converter and that’s why we need a reset winding.
- No core saturation if the peak flux density is chosen properly.
However in Flyback, no mmf balance during ON time therefore cores can reach saturation easily if they are ungapped.
Flyback Transformer (Magnetics)

- Recall Buck-Boost topology, the unisolated version of Flyback. There is a coil for storing energy.
- Flyback transformer is actually a set of magnetic coupled coils.
- Energy is stored in primary winding during the ON time, believe or not, most of the energy is stored in the air gap!
Flyback Transformer (Magnetics)

Turn ratio \( (N_p/N_s) \)

- voltage stress on power switch

\[ V_{\text{stress}} = 1.3 V_{\text{dc max}} + (N_p/N_s)(V_0 + 1) \]

Choose \( V_{\text{stress}} \) with 30\% margin

\[ V_{\text{stress}} = 0.7 V_{\text{breakdown}} \]
Flyback Transformer (Magnetics)

Find $T_{ON\text{max}}$ from

For DCM, dead time of 0.2T

$T_{ON\text{max}} = \frac{(N_p/N_s)(V_o + 1)(0.8 \ T)}{V_{dc} + (N_p/N_s)(V_o + 1)}$

For CCM,

$T_{ON\text{max}} = \frac{(N_p/N_s)(V_o + 1)T}{V_{dc}}$
Flyback Transformer (Magnetics)

Peak current

DCM

\[ I_p = \frac{V_{dc} \cdot T_{ON\text{max}}}{L_p} \]

CCM

\[ I_p = 1.25 \frac{P_{omax}}{V_{dc} \cdot \left(\frac{T_{ON\text{max}}}{T}\right)} \] is half of that of DCM

Assume that efficiency is 0.8
Flyback Transformer (Magnetics)

Primary inductance

DCM \[ P_{\text{omax}} = \frac{(0.8 \times 0.5L_p I_p^2)}{T} \]

\[ L_p = \frac{(V_{dc} T_{ON\text{max}})^2}{(2.5 T P_{\text{omax}})} \]

CCM

\[ L_p = \frac{(V_{dc} T_{ON\text{max}})^2}{(2.5 T P_{\text{omin}})} \]

Inductance at mode boundary

Assume that efficiency is 0.8
Primary & Secondary Windings

Flyback Topology

DCM

- Wire size of primary winding is
  \[ 289 \, I_p \sqrt{T_{ONmax}/T} \text{ in circular mils} \]

- Wire size of secondary winding is
  \[ 289 \, \left( \frac{N_p}{N_s} \right) \left( I_p \sqrt{(T-T_{ONmax})/T} \right) \text{ in circular mils} \]

500/\sqrt{3}
Primary & Secondary Windings

Flyback Topology

CCM

- Wire size of primary winding is
  \[ 500 I_p \sqrt{\frac{T_{ONmax}}{T}} \] in circular mils

- Wire size of secondary winding is
  \[ 500 \left(\frac{N_p}{N_s}\right) \left( I_p \sqrt{\frac{T-T_{ONmax}}{T}} \right) \] in circular mils
Primary & Secondary Windings

**Flyback Topology**

\[ N_p = 1000 \sqrt{L_p / A_{lg}} \text{ Turns} \]

- \( A_{lg} \), the inductance per 1000 turns with air gap

Avoiding saturation,

- \( N_p I_p \) must be less than \( N I_{sat} \) given by the manufacturer at the chosen air gap

Bigger airgap; smaller \( A_{lg} \); higher \( N I_{sat} \)
Primary & Secondary Windings

What if $A_{lg}$ is not given?

- Well, $A_1$, the inductance per 1000 turns of ungapped core is always given
  
  $A_{lg}$ can be estimated as $A_1(l/\mu)/(l_{air} + 1/\mu)$
  
  - $l$ is the magnetic path length
  - $l_{air}$ is the air gap length
  - $\mu$ is the effective core permeability
Primary & Secondary Windings

What if $N I_{sat}$ is not given?

- First find the $B_{sat}$ in Gauss for the chosen material

$NI_{sat}$ can be estimated as $(B_{sat}(l_{air} + 1/\mu))/(0.4 \pi)$

- $l$ is the magnetic path length in cm
- $l_{air}$ is the air gap length in cm
- $\mu$ is the effective core permeability
Primary & Secondary Windings

TDK PE40 PQ59
\( \mu = 2300, l = 8 \text{ cm}, A_l = 10540 \)
Choose \( B_{\text{sat}} = 3000 \text{ G} \)

\[ A_{lg} = A_l \left( \frac{1/\mu}{l_{\text{air}} + 1/\mu} \right) \]

Ungapped \( N I_{\text{sat}} = 8.3 \)

<table>
<thead>
<tr>
<th>Air gap in mils</th>
<th>( A_{lg} )</th>
<th>( N I_{\text{sat}} )</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
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</thead>
<tbody>
<tr>
<td>4</td>
<td>2688</td>
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<td>6</td>
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<td>2752.6</td>
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<td>8</td>
<td>1540</td>
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<td>56</td>
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<td>11</td>
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<td>12</td>
<td>1080</td>
<td>81</td>
<td>81</td>
<td>40</td>
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<td>7083.2</td>
<td>1727.3</td>
<td>787.0</td>
<td>431.8</td>
<td>276.4</td>
</tr>
</tbody>
</table>

\[ NI_{\text{sat}} = \frac{B_{\text{sat}} (l_{\text{air}} + 1/\mu)}{(0.4 \pi)} \]

\( L \) in microhenries

\( N_{\text{max}} \)
\( L_{\text{max}} \)
\( N_{\text{max}} \)
\( L_{\text{max}} \)
Powdered Molypermalloy (MPP) Cores

- Powder is Square Permalloy 80, an alloy of nickel, iron, and molybdenum.

- Powdered particles are mixed with plastic resin and cast in the shape of toroids.
Powdered Molypermalloy (MPP) Cores

- Each particle is encapsulated in a resin envelope that provides the core with distributed air gap
- Behaves like gapped core with less EMI
- For building inductors in power supplies, cores of permeability greater than 125 is rarely chosen because of the saturation caused by DC bias current
Powdered Molypermalloy (MPP) Cores

10% drop in inductance
Powdered Molypermalloy (MPP) Cores

- At 10% drop in inductance, the values of \( H \) in Oersteds for materials having \( \mu \) of 14, 26, 60, 125 are 170, 85, 36, and 19

Maximum ampere-turn can be calculated as

\[
NI = H \frac{l_m}{(0.4 \pi)}
\]

- \( l_m \) is the length of magnetic path in cm

\( N_{\text{max}} \) and \( L_{\text{max}} \) are tabulated in given tables
Powdered Molypermalloy (MPP) Cores

\[ NI = H l_m / (0.4 \pi) \]

<table>
<thead>
<tr>
<th>Core</th>
<th>( \mu )</th>
<th>( A_i )</th>
<th>H</th>
<th>( N_f )</th>
<th>( N_{\text{max}}, L_{\text{max}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>55438</td>
<td>125</td>
<td>281</td>
<td>19</td>
<td>162</td>
<td>162, 81, 54, 32</td>
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<td>135</td>
<td>36</td>
<td>308</td>
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<td></td>
<td>11513.6, 2878.4, 1279.3, 460.5</td>
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</tbody>
</table>

Page A6-8
Transformer Copper Losses

At low frequency
- Copper loss = $(I_{\text{rms}})^2 R_{dc}$
- $R_{dc}$ is length of wire times resistive per unit length

At high frequency, there are losses due to
- Skin effect
- Proximity effect
Transformer Copper Losses

- Both skin and proximity effects arise from eddy currents which are induced by varying, high-frequency, magnetic field.
- Skin effect -- the magnetic field is generated from the current carried by the wire itself
- Proximity effect -- the magnetic field is generated from the current in adjacent wires
Skin Effect

- Skin effect causes current in a wire flow only on the surface of the wire
- Reduction in the effective cross-sectional area
- Increase the resistance at high frequency

At high frequency

- Copper loss = \((I_{\text{rms}})^2 R_{ac}\)
- \(R_{ac}\) can be found from the ratio of \(R_{ac}/R_{dc}\)
Skin Depth

- The thickness at which the current density is decreased to 36% of the value at surface
- Skin depth $d = \sqrt{\frac{1}{\pi f \mu \sigma}}$ in m
  - $f$ is frequency in Hz
  - $\mu$ is permeability in H/m
  - $\sigma$ is the conductivity in S/m
- For copper, $d = 2.1\text{mm}$ at $1\text{kHz}$ and $0.067\text{ mm}$ at $1\text{MHz}$

$1\text{A/mm}^2$  \hspace{1cm}  $0.36\text{A/mm}^2$
Skin Depth

- Iron has a conductivity that is only about 6 times less than that of copper, not too bad indeed!
- Why not use cheap iron for transmission of power?
- Because iron has high $\mu$ value, the skin depth is very small even at low frequency, 0.7 mm at 50Hz
### Table of $R_{ac} / R_{dc}$ (Sine-Wave Currents)

AC/DC Resistance Ratio Due to Skin Effects at Various Frequency (Sine-Wave Currents)

<table>
<thead>
<tr>
<th>Wire No.</th>
<th>Dia. in Mils</th>
<th>25kHz</th>
<th>50kHz</th>
<th>100kHz</th>
<th>200kHz</th>
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<td>12</td>
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<td>1.00</td>
<td>1.00</td>
<td>1.00</td>
<td>1.05</td>
</tr>
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</tbody>
</table>
Table of $R_{ac}/R_{dc}$ (Square-Wave Currents)

AC/DC Resistance Ratio Due to Skin Effects at Various Frequency (Square-Wave Currents)

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<th>Wire No.</th>
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<th>100 kHz</th>
<th>200 kHz</th>
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</tr>
</tbody>
</table>
Skin Effect

- Skin effect is more profound when wire diameter is large but thick wire still yields less loss! Getting worse when frequency is high.
- Use a number of small-diameter wires instead of a single large-diameter wire because the total area in annular skin of the small-diameter wires is bigger.
- Litz wire; take good care in soldering.
Skin Effect

- Skin effect is more profound for square-wave currents because they have a lot of high-frequency harmonics.
Proximity Effect

- Proximity effect is caused by high-frequency magnetic fields arising from currents in adjacent wires or winding layers.

Effective area for current flow
Proximity Effect

- Proximity effect is more profound when the no. of winding layers is large

A 3-layer secondary winding in a EE core
Proximity Effect

Primary/secondary without interweaving

- Higher $R_{ac}/R_{dc}$
- mmf

Diagram:

- Core - Centre leg
- Sec-- Layer 2
- Sec-- Layer 1
- Pri-- Layer 2
- Pri-- Layer 1
Proximity Effect

- Primary/secondary with interweaving

- Lower $\frac{R_{ac}}{R_{dc}}$

- Sec-- Layer 2
- Pri-- Layer 2
- Sec-- Layer 1
- Pri-- Layer 1

- Core centre leg

- mmf