High Frequency Planar Magnetics for Power Conversion

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Outline

- Magnetics Basics Introduction (WGH)
  - Transformer Design
  - High Frequency Effects in the Winding

- Analytical Models for Planar Magnetics (WGH)

- Planar Magnetics – Fundamentals (ZO)

- Planar Magnetic Components Integration (ZO)
Magnetics Basics
Laws of Electromagnetism

Ampere’s Law

\[ \sum H \cdot l = Ni \]

Faraday’s Law

\[ e = -N \frac{d\phi}{dt} \]

\[ B = \mu H \quad \mu = \mu, \mu_0 \]
Losses in Magnetic Components

- Core losses
- Hysteresis loss
- Eddy current loss
- Copper losses
- Skin effect loss
- Proximity effect loss
Core Loss

Hysteresis loss in a ferromagnetic material

Eddy current loss in a ferromagnetic material

\[ P_{fe} = K_v f^\alpha B_{\text{max}}^\beta \]

- Hysteresis loss is the area inside the B-H loop
- Eddy current loss is reduced by laminations
Ferromagnetic Materials

(a) Hard magnetic materials  
(b) Soft magnetic materials
The magnetic and operating properties of some soft magnetic materials

<table>
<thead>
<tr>
<th>Materials</th>
<th>Ferrites</th>
<th>Nanocrystalline</th>
<th>Amorphous</th>
<th>Si Iron</th>
<th>Ni-Fe (Permalloy)</th>
<th>Powdered iron</th>
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</thead>
<tbody>
<tr>
<td><strong>Model</strong></td>
<td>TDK P40</td>
<td>VIROPERM 500F</td>
<td>METGLAS 2605</td>
<td>AK Oriented M-4</td>
<td>MAGNETICS PERMALLOY 80</td>
<td>MICROMET-ALS 35μ</td>
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<tr>
<td>Permeability, $\mu_i$</td>
<td>1500-4000</td>
<td>15000</td>
<td>10,000-150,000</td>
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<td>$B_{peak}, T$</td>
<td>0.45-0.81</td>
<td>1.2</td>
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<td>$\rho, \mu\Omega m$</td>
<td>$6.5 \times 10^6$</td>
<td>1.15</td>
<td>1.3</td>
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<td>$P_{loss}$</td>
<td>60 mW/cm³ at 0.1T/50kHz</td>
<td>588 mW/cm³ at 0.3T/100kHz</td>
<td>72 mW/cm³ at 0.2T/25kHz</td>
<td>2.295-30.6mW/cm³ at 1.5T/50Hz</td>
<td>192.28mW/cm³ at 0.2T/5kHz</td>
<td>126-315mW/cm³ at 0.1T/10kHz</td>
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<tr>
<td>Core Shape</td>
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</table>
Core Loss Density vs Frequency

Power loss density (mW/cm$^3$)

Frequency (kHz)
**Voltage equation**

\[ V_{\text{rms}} = \frac{k}{\tau/T} f N B_{\text{max}} A_m = K_v f N B_{\text{max}} A_m \]

where

\[ k = \frac{V_{\text{rms}}}{<v>} \quad K_v = \frac{k}{\tau} \frac{k}{T} \]

\( K_v = 4.44 \) for a sinewave
\( = 4.00 \) for a squarewave

**Power equation**

\[ \sum VA = K_v f B_{\text{max}} J k_f k_\text{c} A_p \]

**Variables and Definitions**

- \( V_{\text{rms}} \): the rms value of the applied voltage
- \(<v>\): the average value of the applied voltage
- \( K_v \): voltage waveform factor
- \( k \): the form factor
- \( f \): the frequency of the applied voltage
- \( T \): the period of the applied voltage
- \( \tau \): the interval from the point where the flux density is zero to the point where it is at its maximum value
- \( k_f \): the core stacking factor
- \( J_o \): the current density in each winding
Transformer Losses

Winding losses

Total resistive losses

\[ P_{cu} = \sum RI^2 = \rho \sum_{i=1}^{n} \frac{N_i MLT (J_{a_i} A_{wi})^2}{A_{wi}} \]

\[ k_u = \frac{\sum_{i=1}^{n} N_i A_{wi}}{W_a} \]

is window utilization factor

\[ V_w = MLT \times W_a \]

is volume of the windings

\[ P_{cu} = \rho w V_k J^2 \]

Core losses

\[ P_{fe} = V_e K_c f^\alpha B_{max}^\beta \]

\[ A_p = W_a \times A_c \]

Window area x cross-sectional area

Mean Length of a Turn, MLT

Typical layout of a transformer
Winding losses

\[
P_{cu} = \rho_w V k_a \left[ \frac{\sum VA}{K_v f B_{\text{max}} k_f k_u A_p} \right]^2 = \frac{a}{f^2 B_{\text{max}}^2}
\]

Core losses

\[
P_{fe} = V c k f^\alpha B_{\text{max}}^\beta = b f^\alpha B_{\text{max}}^\beta
\]

Total losses

\[
P = \frac{a}{f^2 B_m^2} + b f^\alpha B_m^\beta
\]

At a given operation frequency, \( f = f_{\text{max}} \),

\[
\frac{\partial P}{\partial B_{\text{max}}} = -\frac{2a}{f^2 B_{\text{max}}^3} + \beta b f^\alpha B_{\text{max}}^{\beta-1} = 0
\]

The minimum losses occur when

\[
P_{cu} = \frac{\beta}{2} P_{fe}
\]
Losses Optimization

\[ P = \frac{a}{f^2 B_m^2} + b f^\alpha B_m^\beta \]

\[ B_o \frac{7\beta-12}{12} f^{\frac{7\alpha-12}{12}} = \frac{2^{7/12}}{(\beta + 2)^{2/3}} \left[ \frac{h k_t \Delta T}{(\rho k_w)^{1/12}} \right]^{2/3} \left[ \frac{K_v \sqrt{k_u}}{\sum VA} \right]^{1/6} \]

\[ A_p = \left[ \frac{\beta + 2}{\beta} \frac{1}{k_u \Delta T} \right]^{4/7} \left[ \frac{\sum VA}{K_v f B_o K_\theta} \right]^{8/7} \]

\[ J_o = \sqrt{\frac{\beta}{\beta + 2}} \frac{h k_t}{\rho k_w} \sqrt{\frac{\Delta T}{k_u}} \frac{1}{A_p^{1/8}} = K_\theta \sqrt{\frac{\beta}{\beta + 2}} \sqrt{\frac{\Delta T}{k_u}} \frac{1}{A_p^{1/8}} \]

For \( h = 10 \text{ W/m}^2 \degree \text{C}, k_c = 5.6, k_w = 10, k_t = 40, \rho = 1.72 \times 10^{-8} \, \Omega \cdot \text{m}, K_\theta = 48.224 \times 10^3 \).
Design Methodology

Specifications: \( \sum \text{VA}, K_f, k_u, \Delta T \)

Select Material: \( B_{\text{sat}}, \rho_c, K_c, \alpha, \beta \)

Calculate \( B_o \)

Select \( A_p \)

Calculate \( A_p \)

Select \( A_{p+1} \)

Calculate \( J_o \)

Select Wires

Calculate Copper Loss

Calculate Core Loss

Calculate High Frequency Losses

Calculate Efficiency, \( \eta \)
Push-pull Converter Transformer

Circuit

Waveforms

\[ v_{p1}, v_{p2}, v_s, i_{d1}, i_{d2}, i_o, i_f, i_e \]

Power Electronics Research Centre, NUI Galway
## Design specifications

<p>| | |</p>
<table>
<thead>
<tr>
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</thead>
<tbody>
<tr>
<td><strong>Input</strong></td>
<td>36 → 72 V</td>
</tr>
<tr>
<td><strong>Output</strong></td>
<td>24 V, 12.5 A</td>
</tr>
<tr>
<td><strong>Frequency, ( f )</strong></td>
<td>50 kHz</td>
</tr>
<tr>
<td><strong>Temperature Rise, ( \Delta T )</strong></td>
<td>35 °C</td>
</tr>
<tr>
<td><strong>Ambient Temperature, ( T_a )</strong></td>
<td>45 °C</td>
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</table>

### Core data: EPCOS N67 Mn-Zn

<p>| | |</p>
<table>
<thead>
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</thead>
<tbody>
<tr>
<td>( K_c )</td>
<td>9.12</td>
</tr>
<tr>
<td>( \alpha )</td>
<td>1.24</td>
</tr>
<tr>
<td>( \beta )</td>
<td>2.0</td>
</tr>
<tr>
<td>( B_{sat} )</td>
<td>0.4 T</td>
</tr>
</tbody>
</table>

### Core loss

\[
P_{fe} = K_c f^\alpha B_m^\beta
\]
Push-pull Converter Transformer

Calculations:

(3) VA ratings of the windings

\[ P_o = (24+1) \times 12.5 = 312.5 \text{ W} \]

\[ \Sigma VA = \left( \frac{1}{k_{pp}} \left( \frac{P_o}{2} + \frac{P_o}{2} \right) + \frac{1}{k_{ps}} \left( \frac{P_o}{2} + \frac{P_o}{2} \right) \right) = \left( \sqrt{2} + \sqrt{\frac{1+D}{D}} \right) P_o \]

\[ = \left( \sqrt{2} + \sqrt{\frac{1+0.67}{0.67}} \right) (312.5) = 935 \]

(4) Optimum \( A_p \)

\[ B_o = \frac{[(10)(40)(35)]^{2/3}}{2^{2/3} \left[ (1.72 \times 10^{-8})(10)(0.4) \right]^{1/12} \left[ (5.6)(9.12)(50 \ 000) \right]^{1.24} } \]

\[ \times \left[ \frac{(4.88)(50000)(1.0)(0.4)}{935} \right]^{1/6} = 0.126 \text{ T} \]

The optimum flux density is less than \( B_{sat} \)

\[ A_p = \left[ \frac{\sqrt{2} \times 935}{(4.88)(50000)(0.126)(1.0)(0.4)(48.2 \times 10^3)\sqrt{(0.4)(35)}} \right]^{8/7} \times 10^8 = 2.693 \text{ cm}^4 \]
### ETD44 Core Data

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Value</th>
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<td>$A_c$</td>
<td>1.73 cm²</td>
</tr>
<tr>
<td>$W_a$</td>
<td>2.78 cm²</td>
</tr>
<tr>
<td>$A_p$</td>
<td>4.81 cm⁴</td>
</tr>
<tr>
<td>$V_c$</td>
<td>17.70 cm³</td>
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<tr>
<td>$k_f$</td>
<td>1.0</td>
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<tr>
<td>$k_u$</td>
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<tr>
<td>MLT</td>
<td>7.77 cm</td>
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<tr>
<td>$\rho_{20}$</td>
<td>1.72 $\mu\Omega$-cm</td>
</tr>
<tr>
<td>$\alpha_{20}$</td>
<td>0.00393</td>
</tr>
</tbody>
</table>
Calculations:

(6) Wire size

\[
J_o = K_t \sqrt{\frac{\Delta T}{2k_u}} \frac{1}{\sqrt{A_p}} = (48.2 \times 10^3) \sqrt{\frac{35}{2(0.4)}} \frac{1}{\sqrt{4.81 \times 10^4}} = 2.62 \times 10^6 \text{ A/m}^2
\]

Primary windings:

\[
I_p = \frac{P_o}{2k_{pp} V_p} = \frac{312.5 / 2}{(0.707)(29.5)} = 7.5 \text{ A}
\]

\[
A_w = I_p / J_o = 2.863 \text{ mm}^2
\]

Skin depth at 50 kHz = 0.295 mm

Standard 0.1 × 30 mm copper foil with a dc resistance of 5.8 mΩ/m @ 20°C meets this requirement or a 2 mm diameter wire.
High Frequency Effects in the Windings
Design Issues for High Frequency

- High frequency winding loss
- Core loss: Steinmetz equation, iGSE.
- Parasitic parameters: leakage inductance, stray capacitance

Skin effect

Proximity effect

Fringing effect
Skin Effect Factor

Eddy currents in a circular conductor

Current distribution in a circular conductor

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Current distribution in a circular conductor

\[
\frac{R_{ac}}{R_{dc}} = 1 + \frac{(r_o/\delta)^4}{48 + 0.8(r_o/\delta)^4} \quad r_o/\delta < 2
\]

The ac resistance is proportional to the square root of frequency at very high frequencies.

\[
\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}
\]
Proximity Effect

Transformer cross-section with current density distribution

Proximity effect factor for sinusoidal excitation
Proximity Effect

Transformer cross-section with current density distribution

\[ \frac{R_{ac}}{R_{dc}} = \phi_{prox}(\Delta) = 1 + \frac{5p^2 - 1}{45} \Delta^4 \text{ where } \Delta = \frac{d}{\delta} \]

\[ \delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \]

- As the number of layers \( p \) increase there is a substantial increase in the ac resistance for a given layer thickness \( d \) and frequency \( f \).
A round conductor of diameter $D$ is equivalent to a square conductor of side length

$$d = \sqrt{\frac{\pi}{4}} D$$

The porosity factor

$$\eta = \frac{Nd}{w}$$

The effective conductivity

$$\sigma_w = \eta \sigma$$

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma_w}}$$

Porosity factor for foils and round conductors
Proximity Effect: arbitrary waveform

An arbitrary periodic current waveform may be represented by its Fourier series

\[ i(t) = I_{dc} + \sum_{n=1}^{\infty} \hat{I}_n \cos(n \omega t + \phi_n) \]

The total power loss due to all the harmonics

\[ P = R_{dc} I_{dc}^2 + \sum_{n=1}^{\infty} R_{dc} \phi_{prox}(\Delta_n) I_{n,rms}^2 \]

So

\[ \delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \]

\[ \Delta_n = \frac{d}{\delta} = \sqrt{n} \frac{d}{\delta_o} = \sqrt{n} \Delta \]

\[ R_{eff} = \frac{I_{dc}^2 + \sum_{n=1}^{\infty} \phi_{prox}(\Delta_n) I_{n,rms}^2}{R_{dc} I_{rms}^2} \]

\[ = \frac{I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,rms}^2 + \frac{5p^2-1}{45} \Delta^4 \sum_{n=1}^{\infty} n^2 I_{n,rms}^2}{I_{rms}^2} \]

\[ \phi_{prox}(\Delta_n) = 1 + \frac{5p^2-1}{45} \Delta^4 = 1 + \frac{5p^2-1}{45} n^2 \Delta^4 \]
Proximity Effect

\[ i(t) = I_{dc} + \sum_{n=1}^{\infty} \hat{i}_n \cos(n \omega t + \varphi_n) \]

\[ I_{\text{rms}}^2 = I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,\text{rms}}^2 \]

\[ \frac{R_{\text{eff}}}{R_{dc}} = \frac{I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,\text{rms}}^2 + \frac{5p^2 - 1}{45} \Delta^4 \sum_{n=1}^{\infty} n^2 I_{n,\text{rms}}^2}{I_{\text{rms}}^2} \]

\[ = 1 + \frac{5p^2 - 1}{45} \Delta^4 \left[ \frac{I'_{\text{rms}}}{\omega I_{\text{rms}}} \right]^2 \]

\[ \Delta = \frac{d}{\delta_o} \]

\[ \frac{di(t)}{dt} = I' = -\omega \sum_{n=1}^{\infty} n \hat{i}_n \sin(n \omega t + \varphi_n) \]

\[ I'^2_{\text{rms}} = \omega^2 \sum_{n=1}^{\infty} n^2 I_{n,\text{rms}}^2 \]
The optimum value of $\Delta$

$$\Delta_{opt} = \frac{15}{5p^2 - 1} \sqrt{\frac{\omega I_{rms}}{di/dt}}_{rms}$$

Finally

$$\frac{R_{eff}}{R_{dc}} = 1 + \frac{1}{3} \left( \frac{\Delta}{\Delta_{opt}} \right)^4$$

$$\left( \frac{R_{eff}}{R_{dc}} \right)_{opt} = \frac{4}{3}$$
Optimum Winding Thickness: Pushpull

\[ \Delta_{opt} = \sqrt{\frac{15}{5p^2 - 1}} \sqrt{\frac{\omega I_{rms}}{I'_{rms}}} \]

Skin depth

\[ \delta_o = \frac{66}{\sqrt{f}} = \frac{66}{\sqrt{(50 \times 10^3)}} = 0.295 \, mm \]

Optimum layer \( \Delta \)

\[ \Delta_{opt} = \sqrt{\frac{D - \frac{8t_r}{3T}}{(5p^2 - 1)15}} = \sqrt{\frac{0.67 - \frac{(8)(0.025)}{3}}{[(5)(6)^2 - 1]/15}} = 0.3342 \]

Optimum layer thickness

\[ d_{opt} = \Delta_{opt} \delta_o = (0.3342)(0.295) = 0.1 \, mm \]
Effective ac resistance: foil \( \frac{R_{\text{eff}}}{R_{\text{dc}}} = \frac{4}{3} = 1.3 \)

AC resistance of round conductor

\[
\frac{R_{\text{ac}}}{R_{\text{dc}}} = 0.25 + (0.5)\left(\frac{r_o}{\delta_o}\right) = 0.25 + (0.5)(\frac{1.0}{0.295}) = 1.95
\]

Could replace solid wire with stranded Litz wire
<table>
<thead>
<tr>
<th>Skin effect</th>
<th>Strand-level</th>
<th>Bundle-level</th>
</tr>
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<tbody>
<tr>
<td></td>
<td><img src="image1" alt="Diagram" /></td>
<td><img src="image2" alt="Diagram" /></td>
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</table>

- Litz wire reduces the window utilisation factor, core may be 30% larger for same temperature rise
- Use strands with diameter less than $\delta/4$
- Proximity effect occurs at strand level when wire is twisted
- Twisting cancels proximity effect at bundle level

• Skin effect acts like a solid conductor at the bundle level
• Use strands with diameter less than $\delta/4$

Litz Wire: proximity effect

- Proximity effect occurs at strand level when wire is twisted
- Twisting cancels proximity effect at bundle level

Interleaving the Windings

Current density distribution before interleaving

Current density distribution after interleaving

Current density distribution after interleaving in FEA

\[
\sum J^i = 2\left| t^2 + 2\left(\frac{1}{3}\right)^2 + \left(\frac{2}{3}\right)^2 \right| = 14.4
\]

\[
\sum J^i = 2\left| t^2 + \frac{1}{3} + 2\left(\frac{1}{3}\right) + 2\right| = 4.4
\]
Fringing (Magnetic Field)

Frequency: 100kHz
Core: Magnetics® port core

Gap in the centre leg

Gap in the outer leg
Fringing (Flux)

Gap in the centre leg

Gap in the outer leg
Fringing (Different Frequencies)

Magnetic Field Intensity

Width of conductor: 0.2mm  Core: Magnetics® port core

Frequency 1kHz

Frequency 100kHz
Fringing (Different Frequencies)

Magnetic Flux

Frequency 1kHz

Frequency 100kHz
Fringing (Different Frequencies)

Current Density

Frequency 1kHz

Frequency 100kHz

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Winding Resistance related to the fringing effect, $g=1\text{mm}$

- Distance 1: 0.5mm
- Distance 2: 1.2mm
- Distance 3: 1.9mm
- Distance 4: 2.6mm
- Distance 5: 3.3mm
- Distance 6: 4.0mm
Analytical Models for Planar Magnetics
Advantages

- **Low profile** — planar magnetic components have a lower profile than their wire wound counterparts due to the fabrication process;

- **Automation** — it is difficult to automate the winding of conventional inductors and transformers, the processes used in planar magnetics are based on advanced computer aided manufacturing techniques. Suitable for SMT.

- **High power densities** — planar inductors and transformers are spread out and this gives them a bigger surface-to-volume ratio than conventional components, this enhances the thermal performance;

- **Predictable parasitics** — with planar magnetics, the windings are precise and consistent, yielding magnetic designs with highly controllable and predictable characteristic parameters.
Disadvantages

- **Turns** — the number of turns in planar device tends to be limited by the manufacturing process;

- **Footprint** — larger footprint compared with its conventional counterpart;

- **Capacitance** — interlayer capacitance introduces resonance at high frequencies;

- **Trade-off** — between magnetic core area and winding window area; between the path length versus the mean length of a turn.
A typical planar transformer with an E-I core
Integrated PCB Magnetics

PCB integrated magnetic toroidal transformer
Thick Film Devices: Photoplots

Photoplots of conducting layers

Masks for dielectric layers

Screen generation
Optical photograph of a microsectioned device (scale 30:1) [1] Reproduced with permission from [1]. Copyright 1999 IEEE.

Low Temperature Co-fired Ceramics

<table>
<thead>
<tr>
<th>Green tape</th>
<th>Blanking</th>
<th>Via punching</th>
<th>Via filling</th>
<th>Conductor print</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

LTCC Process flow

Post-processing & Testing → Co-firing → Laminating → Collating
Silicon integrated microinductor: (a) Top view. Reproduced with permission from [1]. Copyright 2008 IEEE, (b) Cross-section. Reproduced with permission from [2]. Copyright 2005 IEEE


Thin Film Devices: Microfabrication

Layer 1: Electroplated bottom core layer

Layer 2: Insulator layer between bottom core and winding

Layer 3: Electroplated copper winding layer

Layer 4: Insulator layer between top core and winding

Layer 5: Electroplating top core layer

Microfabrication process flow for an inductor
## Technology Comparison

<table>
<thead>
<tr>
<th>Technology</th>
<th>Frequency (Typical)</th>
<th>Power (Typical)</th>
<th>Inductance (Typical)</th>
<th>Size (Typical)</th>
</tr>
</thead>
<tbody>
<tr>
<td>PCB magnetics</td>
<td>20 KHz ~ 2 MHz</td>
<td>1 W ~ 5 kW</td>
<td>10 µH ~ 10 mH</td>
<td>100 mm² ~ 100’s cm²</td>
</tr>
<tr>
<td>Thick Film</td>
<td>&lt; 10 MHz</td>
<td>&lt; 10 W</td>
<td>1 µH ~ 1 mH</td>
<td>&lt; 1 cm²</td>
</tr>
<tr>
<td>LTCC</td>
<td>200 KHz ~ 10 MHz</td>
<td>&lt; 10W</td>
<td>1 µH ~ 1 mH</td>
<td>&lt; 1 cm²</td>
</tr>
<tr>
<td>Thin Film</td>
<td>&gt; 10 MHz</td>
<td>&lt; 1W</td>
<td>10’s ~ 100’s nH</td>
<td>&lt; 10mm²</td>
</tr>
</tbody>
</table>
Advantages and Disadvantages

<table>
<thead>
<tr>
<th>Technology</th>
<th>Integration method</th>
<th>Advantages</th>
<th>Disadvantage</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>PCB</strong></td>
<td>Discrete core on laminated structure or integrated core in laminated structure Parallel or sequential process</td>
<td>Low cost Multilayer structure Thick copper High current High inductance</td>
<td>Low resolution (line width 100µm) Relatively low frequency</td>
</tr>
<tr>
<td><strong>Thick Film</strong></td>
<td>Screen printed on sintered ceramic Sequential build up of multiple layers</td>
<td>Low cost</td>
<td>Difficult to form Long process time and low yield due to sequential build-up</td>
</tr>
<tr>
<td><strong>LTCC</strong></td>
<td>Screen printed on green tapes Parallel multilayer and final cofired structure</td>
<td>Parallel layer process High layer counts Module reliability</td>
<td>Cofireability of materials</td>
</tr>
<tr>
<td><strong>Thin Film</strong></td>
<td>Sequential build up of lithographically defined layers</td>
<td>Precision value (line width 5µm) High tolerance High component density High frequency</td>
<td>Low inductance Equipment costly Limited selection on film materials/ material compatibility</td>
</tr>
</tbody>
</table>
Spiral Coil in Air

Circular concentric filaments in air

\[ M = \mu_0 \sqrt{ar} \frac{2}{f} [\left(1 - \frac{f^2}{2}\right)K(f) - E(f)] \]

\[ f = \sqrt{\frac{4ar}{z^2 + (a + r)^2}} \]

\( K(f) \): complete elliptic integrals of the first kind
\( E(f) \): complete elliptic integrals of the second kind
Mutual Inductance of Planar Coils

Planar coils of rectangular cross-section

\[
M_{12} = \frac{\mu_0 \pi}{\ln \left( \frac{r_2}{r_1} \right) \ln \left( \frac{a_2}{a_1} \right)} \int_0^\infty S(kr_2, kr_1)S(ka_2, ka_1)Q(kh_1, kh_2)e^{4iz}dk
\]

\[
S(kx, ky) = \frac{J_0(kx) - J_0(ky)}{k}
\]

\[
Q(kx, ky) = \frac{2}{k^2} \left[ \cosh k \frac{x+y}{2} - \cosh k \frac{x-y}{2} \right] \quad z > \frac{h_1 + h_2}{2}
\]

\[
= \frac{2}{k} \left( h + e^{-kh} - 1 \right) \quad z = 0, x = y = h
\]

Solve with MATLAB

Add a Magnetic Substrate

Planar coils on a magnetic substrate

\[ Z_t = \frac{j \omega \mu_0 T}{h_1 h_2 \ln \left( \frac{r_2}{r_1} \right) \ln \left( \frac{a_2}{a_1} \right)} \int_0^\infty S(k r_2, k r_1) S(k a_2, k a_1) Q(k h_1, k h_2) \lambda(t) e^{-k(d_1 + d_2)} dk \]
Enhancement of inductance with magnetic substrate: (a) as a function of $\mu_r$, (b) as a function of $t$. 

Effect of Permeability and Thickness
Effect of Frequency

Self impedance of a planar coil on a finite substrate: (a) inductance, (b) resistance.

\[ g(\lambda) = \frac{2\lambda(t_1)\lambda(t_2)e^{-2kr}}{1 - \lambda(t_1)\lambda(t_2)e^{-2kr}} \cosh[k(d_2 - d_1)] \]
Sandwich Structure

Planar coils in a sandwich structure

\[ Z_{s\omega}^p = \frac{j \omega \mu_0 \pi}{h_1 h_2 \ln\left(\frac{r_2}{r_1}\right) \ln\left(\frac{a_2}{d_1}\right)} \int_0^\infty S(kr_2, kr_1)S(ka_2, ka_1) \left[ f(\lambda) + g(\lambda) \right] Q(kh_1, kh_2) dk \]

\[ f(\lambda) = \frac{\lambda(t_1)e^{-k(d_1 + d_2)}}{1 - \lambda(t_1)\lambda(t_2)e^{-2ks}} + \frac{\lambda(t_2)e^{-k(d_1' + d_2')}}{1 - \lambda(t_1)\lambda(t_2)e^{-2ks}} \]

\[ g(\lambda) = \frac{2\lambda(t_1)\lambda(t_2)e^{-2ks} \cosh[k(d_2 - d_1)]}{1 - \lambda(t_1)\lambda(t_2)e^{-2ks}} \]
Inductance as a function of substrate separation; $L_1 = 17.14\, \text{nH}$. 

$\frac{L}{L_1} = 1000$ at $s = 0$

$\frac{R}{R_{dc}}$ vs. Frequency (Hz)

- $s = 15\, \mu\text{m}$
- Single substrate
References

Planar Magnetics – Fundamentals
With rapidly increased frequencies, magnetics become **VERY important** factor to achieve high-efficiency and high-power-density converter.
Advantages

- **Low profile**: the height of a planar magnetic core is typically 25% to 50% the height of its wire-wound counterpart.
Advantages

- **Good thermal characteristic:** planar cores essentially have a higher surface area to volume ratio than conventional magnetic cores.

![Finite Element Simulation](image)

(a) Conventional Magnetic Core

(b) Planar Magnetic Core

*Source: DTU Technical Univ. of Denmark*
Advantages

- **Ease of manufacturability and cost reduction:** automation process and computer aided.

- **Modularity:** no extra connections are required.
Advantages

- **Predictable parasitics**: windings manufactured by PCB machines are more precise and consistent, resulting in magnetic designs with highly controllable and predictable parasitic parameters.

- **Ease of implementation on interleaved windings**: multi-layer PCBs allow for an interconnection between arbitrary layers.
In typical planar transformers, most of external flux (leakage flux) is parallel to the surface of the conductors.
Modelling ac resistance in planar transformer is the same with traditional one-dimensional Dowell’s analysis:

\[
\frac{R_{ac}}{R_{dc}} = \varepsilon \left[ \frac{\sinh 2\varepsilon + \sin 2\varepsilon}{\cosh 2\varepsilon - \cos 2\varepsilon} + \frac{2(p^2 - 1)}{3} \frac{\sinh \varepsilon - \sin \varepsilon}{\cosh \varepsilon + \cos \varepsilon} \right]
\]

where \( \varepsilon \) is the ratio of the conductor thickness to the skin depth. \( p \) is the number of layers.

Note: “radial current distribution” in planar structures may affect the dc resistance, but no effect on the ratio of ac resistance to dc resistance.

Fringing Effect

(a) case-1  
Center gap

(b) case-2  
Side gap

(c) case-3  
Horizontal gap

(d) case-4  
Equalized gap

(e) case-5  
Center quasidistributed gap

(f) case-6  
Horizontal quasidistributed gap
Fringing Effect

- **Case 1:** Center gap - Fringing effect on winding loss: Highest, Fringing effect on external magnetic field: Low
- **Case 2:** Side gap - Fringing effect on winding loss: High, Fringing effect on external magnetic field: High
- **Case 3:** Horizontal gap - Fringing effect on winding loss: Medium, Fringing effect on external magnetic field: High
- **Case 4:** Equalized gap - Fringing effect on winding loss: Low, Fringing effect on external magnetic field: Medium
- **Case 5:** Center quasi-distributed gap - Fringing effect on winding loss: Low, Fringing effect on external magnetic field: Low
- **Case 6:** Horizontal quasi-distributed gap - Fringing effect on winding loss: Lowest, Fringing effect on external magnetic field: Medium
Parallel Windings

- High current applications
- Currents may not be equally distributed in the paralleled winding layers

Case 1: [Diagram]
Case 2: [Diagram]
Case 3: [Diagram]
Parallel Windings

At low frequency, “parallel effect losses” or “circulating currents losses” dominates.

At high frequency, eddy current effect losses dominate.

Leakage inductance is simply dependent on the energy stored in core window area:

Leakage energy stored in each elementary layer:

\[ E_{\text{energy}} = \frac{\mu_0}{2} \sum \int_0^h H^2 \cdot l_w \cdot b_w \cdot dx \]
Planar structure is not intrinsically a low leakage inductance construction.

The benefit of planar PCB transformers in this regard is the relative ease with which primary and secondary windings can be heavily interleaved.

Traditional analytical expression:

\[ L_{lk} = \mu_0 \frac{N^2 l_w}{M^2 b_w} \left( \frac{1}{3} \sum_{P=1}^{2M} h_P + \sum_{\Delta=1}^{M} h_{\Delta} \right) \]


layer to layer insulation thickness has not been considered, which in PCB windings (usually have a thicker dielectric layer) may make a significant error.
Leakage Inductance is frequency dependent.


[Ref.2]: J. Zhang, “Analysis and design of high frequency gapped transformers and planar transformers in LLC resonant converters”, PhD thesis, National University of Ireland, 2015.
“radial current distribution” due to high aspect ratio of width to height of a section, $b_w/h_w$. 
The energy stored in the primary/secondary winding is:

\[
E_p = \sum_{i=1}^{n_p} E_i = \frac{\mu_0 \cdot \pi \cdot I_p^2 \cdot n_p \left[ k_1(2n_p^2 + 1) + 4k_2(n_p^2 - 1) \right]}{\ln \left( \frac{r_2}{r_1} \right) \cdot 12 \cdot \gamma \sinh^2(\gamma h_p)}
\]

where,

\[
k_1 = \sinh(2\gamma h_p) - 2\gamma h_p
\]
\[
k_2 = \gamma h_p \cosh(\gamma h_p) - \sinh(\gamma h_p)
\]

The energy stored in the dielectric layer is:

\[
E_d = \frac{1}{2} \cdot \mu_0 \cdot h_i \cdot \int_{r_1}^{r_2} H(r, h)^2 \cdot 2\pi r \cdot dr = \frac{\mu_0 \cdot \pi \cdot h_i}{\ln \left( \frac{r_2}{r_1} \right)} \left[ I_p^2 \sum_{i=1}^{n_p} i^2 + I_s^2 \sum_{i=1}^{n_s-1} i^2 \right]
\]

\[
= \frac{\mu_0 \cdot \pi \cdot h_i}{6\ln \left( \frac{r_2}{r_1} \right)} \left[ I_p^2 \cdot n_p(n_p + 1)(2n_p + 1) + I_s^2 \cdot n_s(n_s - 1)(2n_s - 1) \right]
\]
Leakage Inductance

Giving an example that has only one turn in each layer, the turns ratio \( n = \frac{n_s}{n_p} \) is defined, and all windings’ thickness are the same \((h_p = h_s)\), then the total leakage inductance is:

\[
L_{lk} = \frac{\mu_0 \cdot \pi \cdot n_p}{3 \ln\left(\frac{r_2}{r_1}\right)} \left\{ n_p^2 (k_1 + 2k_2)(n + 1) + \frac{(k_1 - 4k_2)(n + 1)}{2ny\sinh^2(\gamma h_p)} + \left[2(1 + n) \cdot n_p^2 + \frac{1}{n + 1}\right] h_i \right\}
\]

not applicable to complex interleaved cases such as primary and secondary windings on the same layer where 2-D consideration may be needed.
Leakage Inductance

3C96 ER51 cores

Primary winding (8 layers)  Secondary winding (8 layers)

without the consideration of layer to layer insulation

without the consideration of radial current distribution

New model calculation
Leakage Inductance

- Reduce the number of turns. (core saturation and higher core loss)
- Reduce the thickness of conductors and insulators. (high winding resistance and high interwinding capacitance)
- Reduce the mean turn length.
- Increase the window width. (high interwinding capacitance)
- Interleaving winding arrangement.

Small Leakage Inductance is expected in most of power converters

Note: “trade-off”

Leakage Inductance

- Insertion of magnetic shunt (f.x. ferrite polymer composites FPC)

Higher Leakage Inductance is expected in resonant converters such as LLC, DAB etc.

Leakage Inductance

- Thickness of magnetic sheet
- Permeability of magnetic sheet

Leakage Inductance

**MMF method**

**Reluctance method**

\[
L_{tk} = \frac{1}{3} \mu_0 w_c \sum_{i=p}^2 \sum_{i=p}^2 2 n_i \left( t_i + t_{\Delta i} \right) - 3 t_{\Delta i} + \frac{t_{\Delta i}}{n_i} + 6 \mu_t t_{sh} \right) \cdot \frac{w_c}{b_w} - \frac{3 n_1}{4} \mu_t \alpha_R^2 \frac{b_w A_c}{w_c L_{\phi} R_{s2}^2 \left( \frac{R_{c1}}{R_{c2}} + \frac{1}{2} \right) ^2} \right]
\]

Reluctance method provides a better prediction

Higher Leakage Inductance is expected in resonant converters such as LLC, DAB etc.

- **Fractional turn**

(a) $N_p \rightarrow N_s$

(b) $N_p \rightarrow N_s$
\( C_{po}, \ C_{so} \) are self-capacitances of the primary and the secondary windings, respectively.

\( C_{ps0} \) is the mutual capacitance between the two windings.
Winding Capacitance

U-type winding scheme

\[ E = \sum_{i=1}^{n} E_i = \frac{1}{2} C_0 U^2 \sum_{i=1}^{n} \left( \frac{n+1-i}{n} \right)^2 \]

\[ = \frac{(n+1)(2n+1)}{12n} C_0 U^2. \]

\[ C_d = \frac{2E}{U^2} = \frac{(n+1)(2n+1)}{6n} C_0. \]
Winding Capacitance

Z-type winding scheme

\[ C_d = (n/4) \cdot C_o \]
Due to a higher ratio of the width to the thickness of the conductors (intrinsic property of PCB magnetics), $C_1$ is much lower than $C_0$. So, vertical winding scheme leads to a lower electric potential energy.
Interleaved Winding

- Reduce the ac winding resistance; 
  (not for the flyback converter)
- Reduce the leakage inductance;
- Increase the interwinding capacitance.

Interleaved Winding

Leakage energy distribution

Current density distribution

High

Low
Interleaved Winding

- (a) Non-interleaving
- (b) P-S-P-S-P-S-P-S
- (c) P-S-S-P-P-S-S-P
- (d) 0.5P-S-P-S-P-S-P-S-P-S-0.5P

Graphs showing:
- $R_{ac}$ (Ohm) vs Frequency (Hz)
- Leakage Inductance (H) vs Frequency (Hz)
Planar Magnetic Components Integration
**Introduction**

functional devices integration, in which discrete magnetic devices with different functions are assembled as one integrated magnetic device.

Mixing the functions of transformers and inductors in:

- Current doubler rectifiers
- LLC resonant converters
- Integrated EMI filters etc.
Major advantage:

- Smaller size
- Higher power efficiency
- Lower core loss (spark interest into the light load conditions with the integrated magnetics)

It is often used in applications where space is highly restricted such as computer systems, data center, automotive electrical systems and space applications.
● Magnetic core sharing

(Only the core is shared, and the windings are not shared)
Both winding and magnetic core sharing
(Fully integration)
The magnetic field produced by one coil passes through the other coil, changing their effective inductances due to the mutual relationship:

- Coupled Inductors
- Multi-winding transformers
- ...

many advantages,... but can NOT be applied in all circuit topologies
The magnetic field produced by one coil does not pass through the other coil or cancel in the other coil. Can be used in nearly all circuit topologies due to their independent operation behaviors.
Decoupling Methods

- Shared low core reluctance path

- Flux cancellation
Decoupling Methods

- Orthogonal flux path

Four Quadrants Integrated Transformer

Four Quadrants Integrated Transformer

- Wide input range
- Simple control and communication
- High reliability
- Low overall system cost
- Efficient thermal management
- Compact packaging

Magnetic Reluctance Modeling

Applying reluctance-resistance analogy

\[ \begin{align*}
\Phi_1 &= N_1 \cdot I_1 \\
\Phi_2 &= N_2 \cdot I_2 \\
R_1 &= \frac{N_1^2}{R_c} \\
R_2 &= \frac{N_2^2}{R_c}
\end{align*} \]

Reluctance Model

\[ \begin{align*}
\Phi_1 &= \frac{1}{R_c} \\
\Phi_2 &= \frac{1}{R_c}
\end{align*} \]

Inductance Model

For a practical interest that the magnetic and electrical circuits interact
Gyrator-Capacitor Modeling

- Energy interchange between windings and cores
- Understand energy relations and dynamics in the context of power electronics

Coupled inductors

Applying reluctance-model:

Since Kirchhoff's laws and Faraday's law,

\[
\begin{bmatrix}
    v_1 \\
    v_2
\end{bmatrix} = \begin{bmatrix}
    \frac{N_1^2(R_2+R_c)}{\Delta} & \frac{-N_1N_2R_c}{\Delta} \\
    \frac{-N_1N_2R_c}{\Delta} & \frac{N_2^2(R_1+R_c)}{\Delta}
\end{bmatrix} \cdot \begin{bmatrix}
    \frac{di_1}{dt} \\
    \frac{di_2}{dt}
\end{bmatrix} = \begin{bmatrix}
    L_{11} & L_M \\
    L_M & L_{22}
\end{bmatrix} \cdot \begin{bmatrix}
    \frac{di_1}{dt} \\
    \frac{di_2}{dt}
\end{bmatrix}
\]

where,

\[
\Delta = R_1 \cdot R_2 + R_1 \cdot R_c + R_2 \cdot R_c
\]
Coupling coefficient:  (assuming \( N1 = N2 \))

\[
k = \frac{L_M}{L_{11}} = -\frac{N_2 \cdot R_c}{N_1 \cdot (R_2 + R_c)}
\]

- **(a)** \( R_2 \gg R_c \)
  \[ k \approx 0 \]

- **(b)** \( R_2 = 2 \cdot R_c \)
  \[ k \approx \frac{1}{3} \]

- **(c)** \( R_2 \ll R_c \)
  \[ k \approx 1 \]
Coupled inductors

Effective inductances are dependent of circuit operation

Integrated magnetics is actually an “open” technology, and may create “innovation” ideas.
Integrated magnetics has advantages less number of components, smaller size, and potentially higher power efficiency.

Integrated magnetics are not all advantageous. The main issue is to produce unwanted parasitic capacitances among the inductive elements, and a limited power capability.

Must understand the principle of circuit operation.

New geometry core may create something interesting.
Conclusion and Trends
- Planar magnetics still gain its popularity due to low profile and easy manufacture

- Planar magnetics towards very high frequency (so-called micro-inductor/transformer) would be interesting. Accurate 2D/3D models are emergency

- High frequency magnetic materials are emergency

- New winding and core geometries may create something innovative
Thank you very much!

Questions?