Optimal Design of Inductive Components Based on Accurate Loss and Thermal Models

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Introduction

System Layer

Component Layer

Material Layer

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Introduction

Application of Inductive Components (1) : Buck Converter
(DC Current + HF Ripple)

Modeling Difficulties
- Non-sinusoidal current / flux waveform
- Current / flux is DC biased

Solutions
- FFT of current waveform for the calculation of winding losses
- Determine core loss energy for each segment and for each corner point in the piecewise-linear flux waveform
- Loss Map enables to consider a DC bias
Introduction
Application of Inductive Components (2) : Inductor of DAB Converter
(Non-Sinusoidal AC Current)

Schematic

Current / Flux Waveform

Modeling Difficulties
- Non-sinusoidal current / flux waveform
- Core losses occur in the interval of constant flux

Solutions
- FFT of current waveform for the calculation of winding losses
- Improved core loss equation that considers relaxation effects
Introduction
Application of Inductive Components (3) : Three-Phase PFC
(Sinusoidal Current + HF Ripple)

Schematic

Modeling Difficulties
- Non-sinusoidal current / flux waveform
- Major loop and many (DC biased) minor loops

Solutions
- FFT of current waveform for the calculation of winding losses
- Determine core loss energy for each segment and for each corner point in the piecewise-linear flux waveform (-> minor loop losses)
- Add major loop losses
Introduction
Overview About Different Flux Waveforms

- Sinusoidal
- DC Current + HF Ripple
- Non-Sinusoidal AC Current
- Sinusoidal Current + HF Ripple

Minor Loop
Major Loop
Introduction

System Layer

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Material Layer

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Introduction
Overview About Other Modeling Issues
Introduction
Wide Range of Realization Options

Inductors / Transformers

Core Shapes

Conductor Shapes

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Introduction
Modeling Inductive Components (1)

Procedure

1) A reluctance model is introduced to describe the electric / magnetic interface, i.e. $L = f(i)$.

2) Core losses are calculated.

3) Winding losses are calculated.

4) Inductor temperature is calculated.
Introduction
Modeling Inductive Components (2)

The following effects will be taken into consideration:

**Magnetic Circuit Model (e.g. for Inductance Calculation):**
- Air gap stray field
- Non-linearity of core material

**Core Losses:**
- DC Bias
- Different flux waveforms (link to circuit simulator)
- Wide range of flux densities and frequencies
- Different core shapes

**Winding Losses:**
- Skin and proximity effect
- Stray field proximity effect
- Effect of core on magnetic field distribution
- Litz, solid, and foil conductors
Introduction

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Introduction

Motivation for an Accurate Loss Modeling: Multi-Objective Optimization (1)

PFC Rectifier with Input LCL filter

Filter Losses vs. Filter Volume

Converter Losses vs. Converter Volume

Sometime there are parameters that bring advantages for one subsystem while deteriorating another subsystem (e.g. frequency in above example).
In order to get an optimal system design, an overall system optimization has to be performed.

It is (often) not enough to optimize subsystems independent of each other.
Introduction
Motivation for an Accurate Loss Modeling: Multi-Objective Optimization (3)

PFC Rectifier with Input LCL filter

Limits concerning mains
- Tolerable mains harmonics.
- Max. admissible VAr consumption.

Limits concerning rectifier
- Max. admissible $T_{j,max}$
- Max. cooling system vol. $V_{CS,max}$

Limits concerning filter structure
- Max. admissible volume
- Max. admissible losses

Optimize for
- Overall PFC rectifier volume
- Overall PFC rectifier losses
- (PFC system cost)
Outline

- Magnetic Circuit Modeling
  - Core Loss Modeling
  - Winding Loss Modeling
  - Thermal Modeling
  - Multi-Objective Optimization
  - Summary & Conclusion
Magnetic Circuit Modeling
Reluctance Model

Electric Network  Magnetic Network

Conductivity / Permeability  \( \kappa \)  \( \mu \)

Resistance / Reluctance  \( R = l / (\kappa A) \)  \( R_m = l / (\mu A) \)

Voltage / MMF  \( V = \int_{P_2}^{P_1} \vec{E} \, d\vec{s} \)  \( V_m = \int_{P_2}^{P_1} \vec{H} \, d\vec{s} \)

Current / Flux  \( I = \iiint_A \vec{J} \, d\vec{A} \)  \( \Phi = \iiint_A \vec{B} \, d\vec{A} \)
A reluctance model is needed in order to

- calculate the inductance \( L = \frac{N^2}{R_{\text{tot}}} \)
- calculate the saturation current
- calculate the air gap stray field
- calculate the core flux density
Magnetic Circuit Modeling

Core Reluctance

\[
R_m = \frac{l_i}{\mu_0 \mu_r A_i}
\]

Core Reluctance Dimensions

<table>
<thead>
<tr>
<th>Section</th>
<th>( l_i )</th>
<th>( A_i )</th>
</tr>
</thead>
<tbody>
<tr>
<td>I</td>
<td>( b )</td>
<td>( c \cdot t )</td>
</tr>
<tr>
<td>II</td>
<td>( d )</td>
<td>( a \cdot t )</td>
</tr>
<tr>
<td>III</td>
<td>( \frac{2\pi}{4} \cdot \frac{(a+c)}{4} = \frac{\pi}{8} (a + c) \cdot \frac{t(a+c)}{2} )</td>
<td></td>
</tr>
</tbody>
</table>

Mean magnetic length

Mean magnetic cross-sectional area
Magnetic Circuit Modeling
Air Gap Reluctance: Different Approaches (1)

Assumption of Homogeneous Field Distribution

\[ R_m = \frac{I_g}{\mu_0 A_g} \]

- \( I_g \): Air gap length
- \( A_g \): Air gap cross-sectional area

Increase of the Air Gap Cross-Sectional Area

e.g. [1] (for a cross section with dimension \( a \times t \)):

\[ R_m = \frac{I_g}{\mu_0 (a + I_g)(t + I_g)} \]

Magnetic Circuit Modeling
Air Gap Reluctance: Different Approaches (2)

Schwarz-Christoffel Transformation

Transformation Equation $z(t)$

$$z(t) = \frac{1}{\pi} \left(2 \ln \left(1 + \sqrt{1 - t}\right) - \ln t - 2 \sqrt{1 - t}\right)$$

($z$ and $t$ are complex numbers)

Transformation Equation $v(t)$

$$v(t) = \frac{V}{\pi} \ln t$$

Solution to 2-D problems found in literature, e.g. in [3]

Can’t be directly applied to 3-D problems.

Some 3-D solution to problem found in literature; however, they are complex [4] and/or limited to one air gape shape [5]

More simple and universal model desired.


Magnetic Circuit Modeling

Aim of New Model

Air gap reluctance calculation that

- considers the three dimensionality,

- is reasonable easy-to-handle,

- is capable of modeling different shapes of air gaps,

- while still achieving a high accuracy.

Illustration of Different Air Gap Shapes:
Magnetic Circuit Modeling
New Model (1)

Basic Structure for the Air Gap Calculation (2-D) [3]

\[ R'_{\text{basic}} = \frac{1}{\mu_0 \left[ \frac{w}{2l} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi h}{4l} \right) \right]} \]
Magnetic Circuit Modeling
New Model (2)

2-D (1)

Basic Structure for the Air Gap Calculation
Magnetic Circuit Modeling

New Model (3)

2-D (2)

Air Gap Type 1

Air Gap Type 2

Air Gap Type 3

Basic Structure for the Air Gap Calculation

\[ R'_{\text{basic}} = \frac{1}{\mu_0 \left[ \frac{w}{2l} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi h}{4l} \right) \right]} \]
Magnetic Circuit Modeling

New Model (4)

2D $\rightarrow$ 3D : Fringing Factor (1)

Illustrative Example

Air gap per unit length

$\sigma_y = \frac{R'_{zy}}{l_g} \frac{\mu_0 a}{\mu_0 t}$

$\sigma_x = \frac{R'_{zx}}{l_g} \frac{\mu_0 t}{\mu_0 t}$
Magnetic Circuit Modeling
New Model (5)

2D $\rightarrow$ 3D : Fringing Factor (2)

Illustrative Example

3-D Fringing Factor:

$$\sigma = \sigma_x \sigma_y$$

$$R_g = \sigma \frac{l_g}{\mu_0 a t}$$

"Idealized" air gap
(no fringing flux)

Alternative Interpretation:
Increase of air gap cross sectional area

$$\frac{1}{\sigma_x} \cdot A \rightarrow \frac{1}{\sigma_y} \cdot A' \rightarrow A''$$

Magnetic Circuit Modeling
FEM Results

3-D FEM Simulation

Modeled Example

\[ a = 40 \text{ mm}; \ h = 40 \text{ mm} \]

Results

\[ R_g [10^3 \text{A/Vs}] \]

- (homogeneous field distribution)
- (increase of \( A_s \))
- (new)
- FEM

\[ l_s/a \]
Magnetic Circuit Modeling

Inductance Calculation
EPCOS E55/28/21, \( N = 80 \)

**TABLE I**
**MEASUREMENT RESULTS OF E-CORE**

<table>
<thead>
<tr>
<th>Air Gap Length</th>
<th>Calculated classically (3)</th>
<th>Calculated with new approach (12)</th>
<th>Measured</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.0 mm</td>
<td>1.42 mH</td>
<td>1.97 mH</td>
<td>2.07 mH</td>
</tr>
<tr>
<td>1.5 mm</td>
<td>0.96 mH</td>
<td>1.47 mH</td>
<td>1.58 mH</td>
</tr>
<tr>
<td>2.0 mm</td>
<td>0.72 mH</td>
<td>1.22 mH</td>
<td>1.26 mH</td>
</tr>
</tbody>
</table>

Saturation Calculation
EPCOS E55/28/21, \( N = 80 \), \( l_g = 1 \text{ mm}, B_{\text{sat}} = 0.45 \text{ T} \)

Measurement
\( I_{\text{sat}} = 3.7 \text{ A} \)
**Magnetic Circuit Modeling**  
Non-Linearity of the Core Material

**Flux and Reluctance Calculation**

\[ \emptyset = f(R_m(\emptyset), I) \]

This equation must be solved iteratively by using a numerical solving method, e.g. the Newton’s method.

**Reluctance Model**

\[ R_m = f(\emptyset) \quad \emptyset = f(R_m(\emptyset), I) = f(\emptyset, I) \]

**Inductance Calculation**

\[ I = \frac{N^2}{R_{tot}(I)} \]
Example
Introduction

Aim
Design PFC rectifier system.
Show trade-off between losses and volume.
Illustrative example.

Modeling of boost inductors (three individual inductors $L_{2a} = L_{2b} = L_{2c}$) will be step-by-step illustrated in the course of this presentation.
Example
Reluctance Model (1)

Photo & Dimensions

Material
Grain-oriented steel (M165-35S)

Calculation of Core Reluctances

\[ R_{c1} = \frac{d_0 - 2a}{\mu_0 \mu, t} + 2 \frac{\pi (2a)}{\mu_0 \mu, t} \]

\[ = \frac{60 \text{ mm} - 2 \cdot 20 \text{ mm}}{\mu_0 \cdot 20'000 \cdot 20 \text{ mm} \cdot 28 \text{ mm}} + 2 \frac{\pi (2 \cdot 20 \text{ mm})}{\mu_0 \cdot 20'000 \cdot 28 \text{ mm} \cdot (2 \cdot 20 \text{ mm})} = 3654 \frac{A}{Vs} \]

\[ R_{c2} = \frac{d_0 - 2a + 2b}{\mu_0 \mu, t} + 2 \frac{\pi (2a)}{\mu_0 \mu, t} \]

\[ = \frac{60 \text{ mm} - 2 \cdot 20 \text{ mm} + 2 \cdot h}{\mu_0 \cdot 20'000 \cdot 20 \text{ mm} \cdot 28 \text{ mm}} + 2 \frac{\pi (2 \cdot 20 \text{ mm})}{\mu_0 \cdot 20'000 \cdot 28 \text{ mm} \cdot (2 \cdot 20 \text{ mm})} = 12184 \frac{A}{Vs} \]
Example
Reluctance Model (2)

Calculation Air Gap Reluctances

zy-plane

\[ R'_{zy,1} = \frac{1}{\mu_0} \left[ \frac{a}{2L_g} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi a}{4L_g} \right) \right] \]

\[ R'_{zy,2} = \frac{1}{\mu_0} \left[ \frac{a}{2L_g} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi (b + a)}{4L_g} \right) \right] \]

\[ R'_{zy} = \left( R'_{zy,1} + R'_{zy,2} \right) \frac{R'_{zy,3}}{R'_{zy,1} + R'_{zy,2} + R'_{zy,3}} \]

zx-plane

\[ R_{zx,1} = \frac{1}{\mu_0} \left[ \frac{I}{2L_g} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi I}{4L_g} \right) \right] \]

\[ R_{zx,2} = \frac{1}{\mu_0} \left[ \frac{I}{2L_g} + \frac{2}{\pi} \left( 1 + \ln \frac{\pi (b + a)}{4L_g} \right) \right] \]

\[ R_{zx} = \frac{R'_{zx,1} + R'_{zx,2}}{2} \]

\[ \sigma_y = \frac{R'_{zy}}{L_Z} = 0.72 \]

\[ \sigma_x = \frac{R'_{zx}}{L_Z} = 0.84 \]

Basic Reluctance

Material
Grain-oriented steel (M165-35S)

Inductance

\[ L = \frac{N^2}{R_{c1} + R_{c2} + R_{g1} + R_{g2}} = 2.66 \text{ mH} \]

meas. 2.69 mH
Outline

- Magnetic Circuit Modeling
- Core Loss Modeling
- Winding Loss Modeling
- Thermal Modeling
- Multi-Objective Optimization
- Summary & Conclusion
Core Loss Modeling
Overview of Different Core Materials (1)

- Magnetic Materials
  - Soft Magnetic Materials
    - Iron Based Alloys
      - Ferromagnetics
        - Alloys with some amount of Si, Ni, Cr or Co.
        - Low electric resistivity.
        - High saturation flux density.
    - Powder Iron Cores
      - Consist of small iron particles, which are electrically isolated to each other.
  - Hard Magnetic Materials
    - Ferrites
      - Ferromagnetics
        - Ceramic materials, oxide mixtures of iron and Mn, Zn, Ni or Ca.
        - High electric resistivity.
        - Small saturation flux density compared to ferromagnetics.
    - Amorphous Alloys
      - Made of alloys without crystalline order (cf. glasses, liquids).
    - Nanocrystalline Materials
      - Ultra fine grain FeSi, which are embedded in an amorphous minority phase.

Laminated Cores
Laminations are electrically isolated to reduce eddy currents.
Core Loss Modeling
Overview of Different Core Materials (2)

**Selection Criteria**
- Saturation Flux Density
- Power Loss Density (Frequency Range)
- Price
- etc.

<table>
<thead>
<tr>
<th>Core Type</th>
<th>Saturation Flux Density</th>
<th>Power Loss Density</th>
<th>Price</th>
<th>Shapes Available</th>
</tr>
</thead>
<tbody>
<tr>
<td>Amorphous Alloys</td>
<td>High Sat. Flux Density (1.5T)</td>
<td>Low Losses</td>
<td>Low Price</td>
<td>Limited Available</td>
</tr>
<tr>
<td>Nanocrystalline Materials</td>
<td>High Sat. Flux Density (1.1T)</td>
<td>Very Low Losses</td>
<td>Very High Price</td>
<td>Limited Available</td>
</tr>
<tr>
<td>Ferrite</td>
<td>Low Sat. Flux Density (0.45 T)</td>
<td>Low Losses</td>
<td>Low Price</td>
<td>Many Different</td>
</tr>
<tr>
<td>Powder Iron Core</td>
<td>High Sat. Flux Density (1.5 T)</td>
<td>Moderate Losses</td>
<td>Low Price</td>
<td>Many Different</td>
</tr>
<tr>
<td>Laminated Steel Cores</td>
<td>Very High Sat. Flux Density (2.2T)</td>
<td>High Losses</td>
<td>Low Price</td>
<td>Many Different</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

**LF:** $B_{\text{SAT}} = B_{\text{max}}$  
**HF:** $B_{\text{max}}$ is limited by core losses.
Core Loss Modeling
Overview of Different Core Materials (3)

Core Loss Modeling
Physical Origin of Core Losses (1)

Weiss Domains / Domain Walls

- Spontaneous magnetization.
- Material is divided to saturated domains (Weiss domains).
- In case an external field is applied, the domain walls are shifted or the magnetic moments within the domains change their direction. The net magnetization becomes greater than zero.

B-H-Loop

- The flux change is partly irreversible, i.e. energy is dissipated as heat.
- The reason for this are the so called Barkhausen jumps, that lead to local eddy current losses.
- In case the loop is traversed very slowly, these Barkhausen jumps lead to the static hysteresis losses.
Core Loss Modeling
Physical Origin of Core Losses (2)

\textbf{B-H-Loop}

- If the process would be fully reversible, going from \( B_1 \) to \( B_2 \) would store potential energy in the magnetic material that is later released (i.e. the area of the closed loop would be zero).

- Since the process is partly irreversible, the area of the closed loop represents the energy loss per cycle

\[ W = \oint H dB \]
Core Loss Modeling
Classification of Losses (1)

- (Static) hysteresis loss
  - Rate-independent BH Loop.
  - Loss energy per cycle is constant.
  - Irreversible changes each within a small region of the lattice (Barkhausen jumps).
  - These rapid, irreversible changes are produced by relatively strong local fields within the material.

- Eddy current losses

- Residual Losses – Relaxation losses
Core Loss Modeling
Classification of Losses (2)

- (Static) hysteresis loss

- Eddy current losses
  - Depend on material conductivity and core shape.
  - Affect $BH$ loop.

- Residual Losses – Relaxation losses

Measurements
VITROPERM 500F

\[ i \text{[A]} \] vs. \[ t \text{[µs]} \]
Core Loss Modeling
Classification of Losses (3)

- (Static) hysteresis loss
- Eddy current losses
- Residual losses – Relaxation losses
  - Rate-dependent $BH$ Loop.
  - Reestablishment of a thermal equilibrium is governed by relaxation processes.
  - Restricted domain wall motion.
Core Loss Modeling
Typical Flux Waveforms

Sinusoidal

Triangular

Trapezoidal

Combination

e.g. 50/60 Hz isolation transformer

e.g. Buck / Boost converter

e.g. boost inductor of Dual Active Bridge

e.g. Boost inductor in PFC
Core Loss Modeling
Outline of Different Modeling Approaches

Steinmetz Approach

\[ P = k f^\alpha B^\beta \]
- Simple
- Steinmetz parameter are valid only in a limited flux density and frequency range
- DC Bias not considered
- (Only for sinusoidal flux waveforms)

Loss Separation

\[ P = P_{\text{hyst}} + P_{\text{eddy}} + P_{\text{residual}} \]
- Needed parameters often unknown
- Model is widely applicable
- Increases physical understanding of loss mechanisms

Loss Map Approach

(Loss Database)
- Measuring core losses is indispensable to overcome limits of Steinmetz approach

Hysteresis Model

(e.g. Preisach Model, Jiles-Atherton Model)
- Difficult to parameterize
- Increases physical understanding of loss mechanisms
Core Loss Modeling
Overview of Hybrid Modeling Approach

“The best of both worlds” (Steinmetz & Loss Map approach)

Outline of Discussion

- Derivation of the $i^2GSE$. (1)
- How to measure core losses in order to build loss map. (2)
- Use of loss map. (3)
- How to calculate core losses for cores of different shapes? (4)

Loss Map
(Loss Material Database)

\[ P_v = \frac{1}{T} \int_0^T k_i \left( \frac{dB}{dt} \right)^\alpha (\Delta B)^{\beta-\alpha} \, dt + \sum_{l=1}^{n} Q_{cl} P_{cl} \]

(1)

\[ P = \frac{P_v}{V} \]

(4)
Core Loss Modeling
Derivation of the $i^2$GSE – Motivation (1)

Steinmetz Equation SE

$$P_v = k f^\alpha \hat{B}^\beta$$

- Only sinusoidal waveforms ($\rightarrow$ iGSE).

iGSE

$$P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt$$

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} \cos \theta |^\alpha 2^{\beta-\alpha} \, d\theta}$$

- DC bias not considered
- Relaxation effect not considered ($\rightarrow i^2$GSE)
- Steinmetz parameter are valid only for a limited flux density and frequency range

$P_v$: time-average power loss per unit volume
Core Loss Modeling
Derivation of the $i^2$GSE – Motivation (2)

iGSE [8]

$$P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt$$

$$k_i = \frac{k}{(2\pi)^{\alpha-1} \int_0^{2\pi} |\cos \theta|^\alpha 2^{\beta-\alpha} d\theta}$$

How to apply the formula?

For Sinusoidal Waveforms

$$P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} dt = k f^\alpha \left( \frac{\Delta B}{2} \right)^\beta$$

Idea
- Generalized formula that is applicable for different flux waveforms
- Losses depend on $dB/dt$

Core Loss Modeling
Derivation of the i²GSE – Motivation (3)

Waveform

Results

Conclusion

Losses in the phase of constant flux!
Core Loss Modeling
Derivation of the $i^2$GSE – $B$-$H$-Loop

Relaxation Losses

- Rate-dependent $BH$ Loop.
- Reestablishment of a thermal equilibrium is governed by relaxation processes.
- Restricted domain wall motion.

Current Waveform
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 1 (1)

Waveform

Loss Energy per Cycle

Derivation (1)

Relaxation loss energy can be described with

$$E = \Delta E \left( 1 - e^{-\frac{t_1}{\tau}} \right)$$

$\tau$ is independent of operating point.

How to determine $\Delta E$?
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 1 (2)

$\Delta E$ – Measurements

Waveform

Conclusion

$\rightarrow \Delta E$ follows a power function!

$$\Delta E = k_r \left| \frac{d}{dt} B(t) \right|^{\alpha_r} (\Delta B)^{\beta_r}$$
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 1 (3)

Model Part 1

\[ P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta-\alpha} \, dt + \sum_{l=1}^{n} P_{rl} \]

\[ P_{rl} = \frac{1}{T} k_r \left| \frac{d}{dt} B(t) \right|^\alpha \left( \Delta B \right)^\beta \left( 1 - e^{-\frac{t_1}{\tau}} \right) \]
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 2 (1)

Waveform

Power Loss

Explanation

1) For values of $D$ close to 0 or close to 1 a loss underestimation is expected when calculating losses with iGSE (no relaxation losses included).
2) For values of $D$ close to 0.5 the iGSE is expected to be accurate.
3) Adding the relaxation term leads to the upper loss limit, while the iGSE represents the lower loss limit.
4) Losses are expected to be in between the two limits, as has been confirmed with measurements.
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 2 (2)

Waveform

Model Adaption

$$P_v = \frac{1}{T} \int_0^T k_i \frac{dB}{dt}\left(\frac{\Delta B}{\alpha}\right)^{\beta-\alpha} dt + \sum_{l=1}^n \frac{Q_{nl}}{P_l}$$

$Q_{nl}$ should be 1 for $D = 0$

$Q_{nl}$ should be 0 for $D = 0.5$

$Q_{nl}$ should be such that calculation fits a triangular waveform measurement.

$$Q_{nl} = e^{-q_1 \left(\frac{dB(t_+)}{dt} - \frac{dB(t_-)}{dt}\right)} = e^{-q_1 \frac{D}{1-D}}$$
Core Loss Modeling
Derivation of the $i^2$GSE – Model Derivation 2 (3)

Waveform

Power Loss

- $i^2$GSE
- Upper Loss Limits
- Measured Values

$P [W]$ vs $D$
Core Loss Modeling
Derivation of the i²GSE – Summary

The improved-improved Generalized Steinmetz Equation (i²GSE) [9]

\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^{\alpha} (\Delta B)^{\beta-\alpha} \, dt + \sum_{i=1}^n Q_{rl} P_{rl}
\]

with

\[
P_{rl} = \frac{1}{T} k_r \left| \frac{d}{dt} B(t) \right|^{\alpha_r} (\Delta B)^{\beta_r} \left( 1 - e^{-\frac{t_i}{\tau}} \right)
\]

and

\[
Q_{rl} = e^{-q_r \left| \frac{dB(t+)/dt}{dB(t-)dt} \right|}
\]

Core Loss Modeling
Derivation of the $i^2$GSE – Example

$i^2$GSE

Evaluated for each piecewise-linear flux segment

$$P_v = \frac{1}{T} \int_0^T k_i \frac{dB}{dt} \left( \Delta B \right)^{\beta-\alpha} dt + \sum_{l=1}^{n} Q_{nl} P_{tl}$$

Example

Evaluated for each voltage step, i.e. for each corner point in a piecewise-linear flux waveform.

$$\Delta B \begin{cases} \frac{\Delta B}{T/2-t_\gamma} & \text{for } t \geq 0 \text{ and } t < T/2 - t_\gamma \\ 0 & \text{for } t \geq T/2 - t_\gamma \text{ and } t < T/2 \\ -\frac{\Delta B}{T/2-t_\gamma} & \text{for } t \geq T/2 \text{ and } t < T - t_\gamma \\ 0 & \text{for } t \geq T - t_\gamma \text{ and } t < T \end{cases} \quad \frac{dB}{dt} = \begin{cases} \frac{\Delta B}{T/2-t_\gamma} & \text{for } t \geq 0 \text{ and } t < T/2 - t_\gamma \\ 0 & \text{for } t \geq T/2 - t_\gamma \text{ and } t < T/2 \\ -\frac{\Delta B}{T/2-t_\gamma} & \text{for } t \geq T/2 \text{ and } t < T - t_\gamma \\ 0 & \text{for } t \geq T - t_\gamma \text{ and } t < T \end{cases}$$

$$P_v = \frac{T - 2t_\gamma}{T} k_i \left[ \frac{\Delta B}{T/2-t_\gamma} \right]^{\alpha} \left( \Delta B \right)^{\beta-\alpha} + \sum_{l=1}^{2} Q_{nl} P_{tl}$$

with $Q_{r1} = Q_{r2} = 0$

$$P_{r1} = P_{r2} = \frac{1}{T} k_r \left[ \frac{\Delta B}{T/2-t_\gamma} \right]^{\beta_r} \left( \Delta B \right)^{\beta_r} \left( 1 - e^{-\frac{t_\gamma}{T}} \right)$$
Core Loss Modeling
Derivation of the $i^2GSE$ – Conclusion

Remaining Problems

Steinmetz parameter are valid only in a limited flux density and frequency range.

Core Losses vary under DC bias condition.

Modeling relaxation and DC bias effects need parameters that are not given by core material manufacturers.

Measuring core losses is indispensable!
Core Loss Modeling
Overview of Hybrid Modeling Approach

“The best of both worlds” (Steinmetz & Loss Map approach)

Outline of Discussion
- Derivation of the $i^2\text{GSE. (1)}$
- How to measure core losses in order to build loss map. (2)
- Use of loss map. (3)
- How to calculate core losses for cores of different shapes? (4)

\[
P_v = \frac{1}{T} \int_0^T k_i \left( \frac{dB}{dt} \right)^\alpha (\Delta B)^{\beta-\alpha} dt + \sum_{l=1}^n Q_{cl} P_{d,l}
\]  

(1)

\[
P = \frac{P_v}{V}
\]  

(4)
Core Loss Modeling

Core Loss Measurement – Measurement Principle

Waveforms

- Sinusoidal
- Triangular
- Trapezoidal

Excitation System

- Voltage: 0 … 450 V
- Current: 0 … 25 A
- Frequency: 0 … 200 kHz

Loss Extraction

\[ B(t) = \frac{1}{N_2 \cdot A_e} \int_0^t u(\tau) \, d\tau \]
\[ H(t) = \frac{N_1 \cdot i(t)}{l_e} \]

\[ \frac{P[W]}{V[m^3]} = \int f H \, dB \]
Core Loss Modeling
Core Loss Measurement - Overview

System Overview

\[
P = \frac{f \int \frac{H(t)l_e}{N_1 N_2 A_e} dB(t) dt}{f \int \frac{H(t)l_e}{N_1 N_2 A_e} dB(t) dt} = f \int_{B(0)}^{B(T)} H(B) dB = f \Phi H dB
\]
Core Loss Modeling
Overview of Hybrid Modeling Approach

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Outline of Discussion
- Derivation of the \( i^2 \text{GSE} \). (1)
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\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^{\alpha} (\Delta B)^{\beta-\alpha} dt + \sum_{l=1}^{n} Q_{cl} P_{cl} \quad (1)
\]

\[
\frac{P}{V} \quad (4)
\]
Core Loss Modeling
Needed Loss Map Structure

Typical flux waveform

Content of Loss Map

- Relaxation
  \[ \alpha, \beta, \gamma, \tau \]

- B-H-Relation

- LF

- HF
Core Loss Modeling
Minor and Major Loops

Idea

Losses due to Minor and Major Loops are calculated independent of each other and summed up.

Implementation

Actually, it is not considered how the minor loop closes: each piecewise linear segment is modeled as having half the losses of its corresponding closed loop (cf. next slides).

\[
\frac{P}{V} = \left(\frac{P}{V}\right)_{\text{Major}} + \sum \left(\frac{P}{V}\right)_{\text{Piecewise Linear Segment (+ Turning Point)}}
\]
Core Loss Modeling
Hybrid Loss Modeling Approach (1)
Core Loss Modeling
Hybrid Loss Modeling Approach (2)

These equations arise when one evaluates the $i^2$GSE for symmetric triangular waveforms.

Three loss map operating points are required in order to extract the parameters $\alpha$, $\beta$, and $k$ (or $k_i$).

Evaluated for the according corner point in a piecewise-linear flux waveform.

Evaluated for the according piecewise-linear flux segment.

\[ P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^{\alpha} (\Delta B)^{\beta-\alpha} \, dt + \sum_{l=1}^n Q_{il} P_{vl} \]

\[ P = \frac{P_v}{V} \]
Core Loss Modeling

Hybrid Loss Modeling Approach (3)

Interpolation and Extrapolation

\((H_{DC}^*, T^*, \Delta B^*, f^*)\)

- \(H_{DC}\) and \(T\)
- \(\Delta B\) and \(f\)
Core Loss Modeling
Hybrid Loss Modeling Approach (4)

Advantages of Hybrid Approach (Loss Map and i^2GSE):

- Relaxation effects are considered (i^2GSE).

- A good interpolation and extrapolation between premeasured operating points is achieved.

- Loss map provides accurate i^2GSE parameters for a wide frequency and flux density range.

- A DC bias is considered as the loss map stores premeasured operating points at different DC bias levels.
Core Losses
Summary of Loss Density Calculation

Sinusoidal
\[ P = k f^\alpha B^\beta \]

Triangular
\[ P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta - \alpha} dt + \sum_{l=1}^n Q_{\nu} P_{\nu} \]
with \( Q \geq 0, n = 1 \)

Trapezoidal
\[ P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta - \alpha} dt + \sum_{l=1}^n Q_{\nu} P_{\nu} \]
with \( Q = 1, n = 2 \)

Combination
\[ P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha (\Delta B)^{\beta - \alpha} dt + \sum_{l=1}^n Q_{\nu} P_{\nu} \]
with different \( Q \)'s and \( n \gg 0 \).
Core Loss Modeling
Overview of Hybrid Modeling Approach

“The best of both worlds” (Steinmetz & Loss Map approach)

Outline of Discussion

- Derivation of the $i^2$GSE. (1)
- How to measure core losses in order to build loss map. (2)
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- How to calculate core losses for cores of different shapes? (4)

\[
P_v = \frac{1}{T} \int_0^T k_i \left| \frac{dB}{dt} \right|^\alpha \left( \Delta B \right)^{\beta-\alpha} dt + \sum_{l=1}^n Q_{cl} P_{cl} \tag{1}
\]

\[
P = \frac{P}{V} \tag{4}
\]
Core Loss Modeling
Effect of Core Shape

Procedure

1) The flux density in every core section of (approximately) homogenous flux density is calculated.

2) The losses of each section are calculated.

3) The core losses of each section are then summed-up to obtain the total core losses.

Reluctance Model

\[ \phi = f(R_m(\phi), I) = f(\phi, I) \]
Core Loss Modeling
Effective Core Dimensions of Toroid

Motivation for Effective Core Dimensions
Core loss densities are needed to model core losses. It is difficult to determine these loss densities from a toroid, since the flux density is not distributed homogeneously in a toroid.

Definition: Ideal Toroid
A toroid is ideal when he has a homogenous flux density distribution over the radius \((r_1 \approx r_2)\).

Idea for Real Toroid
Find effective core magnetic length and cross section, so one can calculate as if it were an ideal toroid, i.e. as if the flux density distribution were homogenous.

 Effective magnetic length
\[
I_c = \frac{2\pi \ln \frac{r_2}{r_1}}{\frac{1}{r_1} - \frac{1}{r_2}}
\]

 Effective magnetic cross-section
\[
A_c = \frac{h \ln^2 \frac{r_2}{r_1}}{\frac{1}{r_1} - \frac{1}{r_2}}
\]
Core Loss Modeling
Impact of Core Shape on Eddy Current Losses

Eddy current loss density can be determined as [5]

\[ P_{\text{eddy}} = \frac{(\pi Bfd)^2}{k_{\text{ec}} \rho} \]

For a laminated core it is

\[ P_{\text{eddy}} = \frac{(\pi Bfd)^2}{6 \rho} \]

→ The eddy current losses per unit volume depend not on the shape of the bulk material, but on the size and geometry of the insulated regions.

→ In case of laminated iron cores, it is still appropriate to calculate with core loss densities that have been measured on a sample core with a geometrically different bulk material, but with the same lamination or tape thickness.

Core Loss Modeling
Effect in Tape Wound Cores

Thin ribbons (approx. 20 μm)
Wound as toroid or as double C core.
Amorphous or nanocrystalline materials.

Losses in gapped tape wound cores higher than expected!
Core Loss Modeling
Effect in Tape Wound Cores - Cause 1: Interlamination Short Circuits

Machining process
Surface short circuits introduced by machining
(particular a problem in in-house production).

After treatment may reduce this effect. At ETH, a core was put in an 40% ferric chloride FeCl$_3$ solution after cutting, which substantially (more than 50%) decreased the core losses.
Core Loss Modeling
Effect in Tape Wound Cores - Cause 2: Orthogonal Flux Lines (1)

A flux orthogonal to the ribbons leads to very high eddy current losses!
Core Loss Modeling
Effect in Tape Wound Cores - Cause 2: Orthogonal Flux Lines (2)

An experiment that illustrates well the loss increase due to an orthogonal flux is given here.

Displacements

Horizontal Displacement  Vertical Displacement

Core Loss Results

![Core Loss Results Graph]

- Horizontal displacement
- Vertical displacement

[Misalignment [% of the corresponding dimension] vs. Core Losses [W]]
Core Loss Modeling
Effect in Tape Wound Cores - Cause 2: Orthogonal Flux Lines (3)

Core loss increase due to leakage flux in transformers.

Measurement Set Up

Results

A higher load current leads to higher orthogonal flux!
Core Loss Modeling
Effect in Tape Wound Cores - Cause 2: Orthogonal Flux Lines (4)

In [10] a core loss increase with increasing air gap length has been observed.

Fig. 1 Core loss per cycle W/f in FINEMET, Fe-based amorphous, and ferrite cut cores as a function of inverse of the effective permeability $\mu_r$.

Fig. 2 Schematic representation of in-plane eddy current generated by leakage flux normal to ribbon surfaces.

Example
Core Loss Modeling

Photo & Dimensions

An approximately homogeneous flux density distribution inside the core.

Flux Density Distribution

Reluctance Model

Material
Grain-oriented steel (M165-35S)

Flux Density Waveform

MATLAB Presentation
Outline

- Magnetic Circuit Modeling
- Core Loss Modeling
- Winding Loss Modeling
- Thermal Modeling
- Multi-Objective Optimization
- Summary & Conclusion
Winding Loss Modeling

Skin Effect (1)

**H-field in conductor**

**Ampere’s Law**

\[ \oint H \, dl = \iint J \, dA \]

**Faraday’s Law**

\[ \oint E \, dl = - \frac{d}{dt} \iint B \, dA \]

**Induced Eddy Currents**
Winding Loss Modeling
Skin Effect (2)

FEM Simulation

Current Distribution

Outer radius of conductor
Winding Loss Modeling

Skin Effect (3)

Skin Depth \( \delta \), where the current density has \( 1/e \) of surface value:

\[
\delta = \frac{1}{\sqrt{\pi \mu_0 \sigma f}}
\]

Power Loss Increase with Frequency:

\[
P_S = F_R(f) \cdot R_{DC} \cdot \hat{I}^2
\]
Winding Loss Modeling
Skin Effect (4)

Current Distributions

Parallel-connected (not twisted) conductors!

Figure from [19]
Winding Loss Modeling
Proximity Effect (1)

$H$-field of neighboring conductor induces eddy currents

Ampere’s Law

$$\oint H \, dl = \iint J \, dA$$

Faraday’s Law

$$\oint E \, dl = - \frac{d}{dt} \iint B \, dA$$
Winding Loss Modeling
Proximity Effect (2)

Eddy Currents in Conductor

Induced Eddy Currents

Current Concentration

Figures from [19]

\( H_{e,\text{rms}} = 35 \text{ A/m parallel to conductor} \)
Winding Loss Modeling
Skin vs. Proximity Effect

Situation

\(f = 100 \text{ kHz}, \ I_{\text{peak}} = 1 \text{ A}, \ H_{\text{e,peak}} = 1000 \text{ A/m}\)

Results

<table>
<thead>
<tr>
<th>Definition</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Skin Effect Losses</strong> (P_{\text{Skin}})</td>
<td>Losses due to current (I), including loss increase due to self-induced eddy currents.</td>
</tr>
<tr>
<td><strong>Proximity Effect Losses</strong> (P_{\text{Prox}})</td>
<td>Losses due to eddy currents induced by external magnetic field (H_{\text{e}}).</td>
</tr>
</tbody>
</table>
Winding Loss Modeling

Litz Wire (1) - What are Litz wires?

Idea

Advantages of Litz wires
- HF losses can be reduced substantially

Disadvantages of Litz wires
- High price
- Heat dissipation difficult

Implementation
Winding Loss Modeling
Litz Wire (2) - Why Litz Wires Have to be Twisted? (1)

**Bundle-Level Skin Effect**

![Diagram of Litz Wire Bundle-Level Skin Effect]

- **Induced electric field**
- **Ampere’s Law**

\[ \oint H \, dl = \iint J \, dA \]  
\[ \oint E \, dl = -\frac{d}{dt} \iint B \, dA \]  

**Current Distributions**

![Diagram of Current Distributions]

- **Internal Proximity Effect**
  (will be explained later)
Winding Loss Modeling
Litz Wire (3) - Why Litz Wires Have to be Twisted? (2)

Bundle-Level Proximity Effect

\[ \oint \mathbf{H} \, dl = \iint \mathbf{J} \, dA \quad \text{(Ampere's Law)} \]

\[ \oint \mathbf{E} \, dl = -\frac{d}{dt} \iint \mathbf{B} \, dA \quad \text{(Faraday's Law)} \]

Figures from [19]
Winding Loss Modeling
Litz Wire (4) – Strand-Level Effects

Internal and External Fields lead to Internal and External Proximity Effects

Losses in Litz Wires
Losses in Solid Wires

Skin effect
Internal prox. effect
Ext. prox. effect

(25 \times d_i = 0.5 \text{ mm}, \ I_{\text{peak}} = 5 \text{ A}, \ H_{\text{e,peak}} = 300 \text{ A/m})

(d = 2.5 \text{ mm}, \ I_{\text{peak}} = 5 \text{ A}, \ H_{\text{e,peak}} = 300 \text{ A/m})
Winding Loss Modeling
Litz Wire (5) – Types of Eddy-Current Effects in Litz Wire

Winding Loss Modeling
Litz Wire (6) – Real Litz Wire

How do “real” Litz wires behave? [12]

Skin Effect / Internal Proximity Effect
\[ R_{\text{skin},\lambda} = \lambda_{\text{skin}} R_{\text{skin,ideal}} + (1 - \lambda_{\text{skin}}) R_{\text{skin,parallel}} \]

External Proximity Effect
\[ R_{\text{prox},\lambda} = \lambda_{\text{prox}} R_{\text{prox,ideal}} + (1 - \lambda_{\text{prox}}) R_{\text{prox,parallel}} \]

Litz Wire Type 1: 7 bundles with 35 strands each¹): \( \lambda_{\text{skin}} \approx 0.5 / \lambda_{\text{prox}} \approx 0.99 \)
Litz Wire Type 2: 4 bundles with 61/62 strands each¹): \( \lambda_{\text{skin}} \approx 0.9 / \lambda_{\text{prox}} \approx 0.99 \)

Operating Point
\( f = 20 \text{ kHz} / n = 130 / d_1 = 0.4 \text{ mm} \)

Winding Loss Modeling
Litz Wire (7) - Are Litz Wires Better than Solid Conductors?

Skin and Internal Proximity Effect

- Solid Wire
  - \( d = 2.5 \text{ mm}, I_{\text{peak}} = 5 \text{ A} \)

- Litz Wire
  - \( 25 \times d_i = 0.5 \text{ mm}, I_{\text{peak}} = 5 \text{ A} \)

- Litz Wire
  - \( 50 \times d_i = 0.35 \text{ mm}, I_{\text{peak}} = 5 \text{ A} \)

- Litz Wire
  - \( 100 \times d_i = 0.25 \text{ mm}, I_{\text{peak}} = 5 \text{ A} \)

External Proximity Effect

- Litz Wire of 25 strands with \( d = 0.5 \text{ mm} \)

- Solid Wire with \( d = 2.5 \text{ mm} \)
Winding Loss Modeling
Foil Windings Enclosed by Magnetic Material

Advantages of foil windings
- HF losses can be reduced
- Lower price compared to Litz wire
- High filling factor

Disadvantages of foil windings
- Increased winding capacitance
- Risk of orthogonal flux

"Skin" of foil conductor larger than of round conductor with same cross section; hence, skin effect losses lower in foil conductor.
Winding Loss Modeling
Foil Windings Not Enclosed by Magnetic Material (1)

Orthogonal flux leads to increased skin and proximity effect.
Winding Loss Modeling
Foil Windings Not Enclosed by Magnetic Material (2)

(Foil) Windings with Return Conductors

\[ P \text{ [mW]} \]

\[ f \text{ [kHz]} \]

- \( I_{\text{peak}} = 1 \text{ A} \)
- \( d = 1.95 \text{ mm} \)
- \( 10 \text{ mm} \times 0.3 \text{ mm} \)

FEM @ 8 kHz
Winding Loss Modeling
Overview About Different Winding Types

<table>
<thead>
<tr>
<th>Type</th>
<th>Price</th>
<th>Skin &amp; Proximity</th>
<th>Filling Factor</th>
<th>Heat Dissipation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Round Solid Wire</td>
<td>++</td>
<td>--</td>
<td>+</td>
<td>+</td>
</tr>
<tr>
<td>Litz Wire</td>
<td>--</td>
<td>++</td>
<td>-</td>
<td>--</td>
</tr>
<tr>
<td>Foil Winding</td>
<td>+</td>
<td>+</td>
<td>++</td>
<td>+</td>
</tr>
<tr>
<td>Rectangular Wire</td>
<td>++</td>
<td>-</td>
<td>++</td>
<td>+</td>
</tr>
</tbody>
</table>

Table and Figure from [13] M. Albach, “Induktive Komponenten in der Leistungselektronik”, VDE Fachtagung - ETG Fachbereich Q1 "Leistungselektronik und Systemintegration", Bad Nauheim, 14.04.2011

Foil windings are better only in this region!

\[ P_0 \ldots \text{round conductor} \]
Winding Loss Modeling
Skin Effect of Foil Conductor

Geometry Considered

\[ P_S = F_F(f) \cdot R_{DC} \cdot \hat{I}^2 \]
(Loss per unit length)

with

\[ F_F = \frac{\nu \sinh \nu + \sin \nu}{4 \cosh \nu - \cos \nu} \]
\[ R_{DC} = \frac{1}{\sigma b h} \]
\[ \nu = \frac{h}{\delta} \]
\[ \delta = \frac{1}{\sqrt{\pi \mu_0 \sigma f}} \]

Current Distribution

For the evaluation of \( F_F \):

Current density [A/mm²]

- \( f = 400 \text{kHz} \)
- \( f = 100 \text{kHz} \)
- \( f = 10 \text{kHz} \)
Winding Loss Modeling
Proximity Effect of Foil Conductor

Geometry Considered

\[ P_p = G_F(f) \cdot R_{DC} \cdot \hat{H}^2 \]

(Loss per unit length)

with

\[ G_F = b^2 \cdot \frac{\sinh \nu - \sin \nu}{\cosh \nu + \cos \nu} \]

\[ R_{DC} = \frac{1}{\sigma b h} \]

\[ \nu = \frac{h}{\delta} \]

\[ \delta = \frac{1}{\sqrt{\pi \mu_0 \sigma f}} \]

Current Distribution

\[ \text{Current density [A/mm}^2] \]

\[ f=400\text{kHz} \]

\[ f=100\text{kHz} \]

\[ f=10\text{kHz} \]

\[ (b = 1 \text{ m}) \]

\[ G_F \text{ evaluated} \]
Winding Loss Modeling
Skin Effect of Solid Round Conductor

\[ P_S = F_R \left( f \right) \cdot R_{DC} \cdot \dot{I}^2 \]
(Loss per unit length)

with

\[ R_{DC} = \frac{4}{\sigma \pi d^2} \]
\[ \xi = \frac{d}{\sqrt{2\delta}} \]
\[ \delta = \frac{1}{\sqrt{\pi \mu_0 \sigma f}} \]
\[ F_R = \frac{\xi}{4\sqrt{2}} \left[ \frac{\text{ber}_0(\xi)\text{bei}_1(\xi) - 2}{\text{ber}_1(\xi)^2 + \text{bei}_1(\xi)^2} - \frac{\text{bei}_0(\xi)\text{ber}_1(\xi) - 2}{\text{ber}_1(\xi)^2 + \text{bei}_1(\xi)^2} \right] \]
Winding Loss Modeling
Proximity Effect of Solid Round Conductor

\[ P_p = G_R(f) \cdot R_{DC} \cdot \hat{H}_S^2 \]
(Loss per unit length)

with

\[ R_{DC} = \frac{4}{\sigma \pi d^2} \]
\[ \xi = \frac{d}{\sqrt{2} \delta} \]
\[ \delta = \frac{1}{\sqrt{\pi \mu_0 \sigma f}} \]

\[ G_R = -\frac{\xi^2 d^2}{2 \sqrt{2}} \left[ \frac{\text{ber}_2(\xi) \text{bei}_1(\xi) + \text{ber}_2(\xi) \text{bei}_1(\xi)}{\text{ber}_0(\xi)^2 + \text{bei}_0(\xi)^2} + \frac{\text{bei}_2(\xi) \text{bei}_1(\xi) - \text{bei}_2(\xi) \text{ber}_1(\xi)}{\text{ber}_0(\xi)^2 + \text{bei}_0(\xi)^2} \right] \]
Winding Loss Modeling
Skin and Proximity Effect of Litz Wire

Skin Effect

\[ P_s = n \cdot R_{DC} \cdot F_R(f) \cdot \left( \frac{\hat{i}}{n} \right)^2 \]

(Loss per unit length)

Proximity Effect

\[ P_p = P_{pe} + P_{pi} \]
\[ = n \cdot R_{DC} \cdot G_R(f) \cdot \left( H_e^2 + \frac{\hat{i}^2}{2\pi^2 d_a^2} \right) \]

(Loss per unit length)

Average internal field \( H_i \) under the assumption of a homogeneous current distribution inside the Litz wire.

Losses in Litz Wires

(25 x \( d_i = 0.5 \) mm, \( I_{peak} = 5 \) A, \( H_{e,peak} = 300 \) A/m)
Winding Loss Modeling
Orthogonality of Winding Losses

It is valid to calculate the losses for each frequency component independently and total them up.

\[ P = \sum_{i=0}^{\infty} \left( P_{S,i} + P_{P,i} \right) \]

It is valid to calculate the skin and proximity losses independently and total them up.

Winding Loss Modeling
Calculation of External Field $H_e$ (1D - Approach)

Un-Gapped Transformer Cores

\[
\begin{align*}
P &= R_{DC} \left( F_{R/F} \hat{I}^2 N M + N G_{R/F} \sum_{m=1}^{M} \hat{H}_{avg,m}^2 \right) l_m \\
&= R_{DC} \hat{I}^2 \left( F_{R/F} N M + N^3 M G_{R/F} \frac{4M^2 - 1}{12b_F^2} \right) l_m
\end{align*}
\]

with

\[
H_{avg} = \frac{1}{2} \left( H_{left} + H_{right} \right)
\]

it is

\[
P = R_{DC} \hat{I}^2 \left( F_{R/F} N M + N^3 M G_{R/F} \frac{4M^2 - 1}{12b_F^2} \right) l_m
\]

where

$N$ ... the number of conductors per layer
(i.e. $N = 1$ for foil windings)

$M$ ... the number of layers.
Winding Loss Modeling
Short Foil Conductors

“Porosity Factor”

\[ \eta = \frac{Nb_i}{b_F} \]

Redefinition of Parameters

\[ \sigma' = \eta \sigma \]

\[ \delta' = \frac{1}{\sqrt{\pi f \sigma' \mu_0}} \]

\[ \nu' = \frac{h}{\delta'} \]
Winding Loss Modeling
FEM Simulations: Foil Windings

\[ N = 2 \times 10 \]
\[ h = 0.3 \text{ mm} \]
\[ b_L = 33 \text{ mm} \]
\[ b_F = 37 \text{ mm} \]
\[ I_{\text{peak}} = 1 \text{ A} \]

Error < 6.5%
Winding Loss Modeling
Calculation of External Field $H_e$ (2D - Approach)

Gapped cores: 2D approach is necessary!
Winding Loss Modeling
Effect of the Air Gap Fringing Field

The air gap is replaced by a fictitious current, which …

… has the value equal to the magneto-motive force (mmf) across the air gap.

→ An accurate air gap reluctance model is needed!
Winding Loss Modeling
Effect of the Core Material

The **method of images (mirroring)**

“Pushing the walls away”
Winding Loss Modeling
Calculation of External Field $H_e$ (2D - Approach)

Gapped cores: 2D approach

Winding Arrangement

External field vector across conductor $q_{x_i,y_k}$

$$
\hat{H}_e = \sum_{u=1}^{m} \sum_{l=1}^{n} \epsilon(u,l) \frac{\hat{i}_{x_u,y_l} \left( (y_l - y_k) - j(x_u - x_i) \right)}{2\pi \left( (x_u - x_i)^2 + (y_l - y_k)^2 \right)}
$$
Winding Loss Modeling

Different Winding Sections

Section 1
Many mirroring steps necessary in order to push the walls away.

Section 2
Only one mirroring step necessary (only one wall).

⇒ Normally, higher proximity losses in Section 1.
Winding Loss Modeling

FEM Simulations: Round Windings (Including Litz Wire Windings) (1)

Major Simplification

- Magnetic field of the induced eddy currents neglected.
- This can be problematic at frequencies above (rule-of-thumb) [15]

\[ f_{\text{max}} = \frac{2.56}{\pi \mu_0 \sigma d^2} \]

Results of considered winding arrangements

<table>
<thead>
<tr>
<th>f-range</th>
<th>( f &lt; f_{\text{max}} )</th>
<th>( f &gt; f_{\text{max}} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Error</td>
<td>&lt; 5%</td>
<td>&gt; 5% (always &lt; 25%)</td>
</tr>
</tbody>
</table>

Winding Loss Modeling
FEM Simulations: Round Windings (Including Litz Wire Windings) (2)
Winding Loss Modeling
Methods to Decrease Winding Losses (1)

Interleaving

Optimal Solid Wire Thickness

Optimal Foil Thickness

Litz Wire

Avoid Orthogonal Flux in Foil Windings

(f = 100 kHz, \( I_{\text{peak}} = 1 \text{ A} \), \( H_{e,\text{peak}} = 1000 \text{ A/m} \))

Push this point to higher frequencies!

→ Increase number of strands.
Winding Loss Modeling
Methods to Decrease Winding Losses (2)

Arrangement of Windings

Proximity losses increase in more compact winding arrangements.
Winding Loss Modeling
Methods to Decrease Winding Losses (3)

Aluminum vs. Copper [13]

Aluminum (vs. Copper):
- Lighter
- Lower costs
- Lower Conductivity $\sigma = 38 \cdot 10^6 \ 1/(\Omega m)$ (Copper: $\sigma = 58 \cdot 10^6 \ 1/(\Omega m)$)

$\rightarrow$ Lower Skin Depth!

Skin- and DC losses higher than in copper conductors.

Proximity losses are lower in aluminum conductors over a wide frequency range. Figure shows a comparison of single round solid conductors in external field.

Example
Winding Loss Modeling

Photo & Dimensions

Material
Grain-oriented steel (M165-35S)

Current Waveform

Demonstration in MATLAB
Outline

- Magnetic Circuit Modeling
- Core Loss Modeling
- Winding Loss Modeling
- Thermal Modeling
- Multi-Objective Optimization
- Summary & Conclusion
Thermal Modeling
Motivation & Model (1)

Motivation
Thermal modeling is important to …
… avoid overheating.
… improve loss calculation (since the losses depend on temperature).
Thermal Modeling
Motivation & Model (2)

Model

→ Determination of thermal resistors is challenging!
**Thermal Modeling**

**Heat Transfer Mechanisms**

\[ R_{th} = \frac{\Delta T}{P} = f(T) \]

**Conduction**
Independent of temperature \( T \) for most materials
Difficult to determine interfaces between materials

\[ R_{th} = \frac{\Delta T}{P} = \frac{l}{A\lambda} \]

**Convection**
Combined effect of conduction and fluid flow
Changes with changing absolute temperature (nonlinear)
Good empirical calculation approach available

\[ R_{th} = \frac{\Delta T}{P} = \frac{1}{\alpha A} \]

**Radiation**
Small compared to other mechanisms
Modeling the system is demanding
(nonlinear eq. / to describe which components “sees” the other component).

\[ P = \varepsilon_{eff} A_1 \sigma (T_b^4 - T_a^4) \]
Thermal Modeling
Thermal Resistance Calculation: (Natural) Convection (1)

\[ R_{th} = \frac{\Delta T}{P} = \frac{1}{\alpha A} \]

\( \alpha \) is a coefficient that is influenced by...

... the absolute temperature,
... the fluid property,
... the flow rate of the fluid,
... the dimensions of the considered surface,
... orientation of the considered surface,
... and the surface texture.
Thermal Modeling
Thermal Resistance Calculation: (Natural) Convection (2)

Empirical solutions known for …

vertical plane

horizontal plane
- top:

- bottom:

gap
- horizontal:

- vertical:

and more …
Thermal Modeling

Thermal Resistance Calculation: (Natural) Convection (3)

Structure of Empirical Solutions - Theory

<table>
<thead>
<tr>
<th>Name</th>
<th>Measure of …</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nusselt number</td>
<td>… improvement of heat transfer compared to the case with hypothetical static fluid.</td>
</tr>
<tr>
<td>$Nu$</td>
<td></td>
</tr>
<tr>
<td>Grashof number</td>
<td>… ratio between buoyancy and frictional force of fluid.</td>
</tr>
<tr>
<td>$Gr$</td>
<td></td>
</tr>
<tr>
<td>Prandtl number</td>
<td>… ratio between viscosity and heat conductivity of fluid.</td>
</tr>
<tr>
<td>$Pr$</td>
<td></td>
</tr>
<tr>
<td>Rayleigh number</td>
<td>… flow condition (laminar or turbulent) of fluid.</td>
</tr>
<tr>
<td>$Ra$</td>
<td></td>
</tr>
</tbody>
</table>

Procedure

$Nu(Gr, Pr) \quad \alpha = \frac{Nu(Gr, Pr) \lambda}{l} \quad \Rightarrow \quad R_{th} = \frac{1}{\alpha A}$

$\lambda$ is the heat conductivity of the fluid

$\lambda_{air} = 25.873 \text{ mW/(m K)} @ 20\degree C$ [16]

**Thermal Modeling**

**Thermal Resistance Calculation: (Natural) Convection (4)**

**Structure of Empirical Solutions - Example**

\[ Nu = \left(0.825 + 0.387(Ra \cdot f_1(Pr))^{1/6}\right)^2 \]

\[ f_1 = \left(1 + \left(\frac{0.492}{Pr}\right)^{9/16}\right)^{-16/9} \]

**Reference:**


**Example**

\( h = 10 \text{ cm}, \ T_p = 60 \, ^\circ\text{C}, \ T_a = 20 \, ^\circ\text{C}, \ A = h \cdot h \) \n
\( R_{th} = 16.6 \text{ K/W} \)

**Increase of Winding Surface**

\[ A_2 = A_1 \cdot \pi/2 \]
Thermal Modeling
Thermal Resistance Calculation: Conduction

Overview

Normally, with natural convection it is

$$R_{\text{th, W}} \ll R_{\text{th, WA}}$$

Round Conductor

$$R_{\text{th, W}}$$ is difficult to determine. One difficulty is, e.g., to model the influence of pressure on the thermal resistance.

Foil Conductor

Litz wire shows low thermal conductivity.
Example
Thermal Modeling (1)

(1) Horizontal Plane - Top

\[ Gr = \frac{g l^3}{v^3} \beta \Delta T \]

\[ Pr = 0.7 \quad \text{(for air)} \]

\[ Ra = Pr \cdot Gr \]

\[ Nu = \begin{cases} 0.766 (Ra \cdot f_2(Pr))^{1/5} & \text{for } Ra \cdot f_2(Pr) \leq 7 \cdot 10^4 \quad \text{(laminar)} \\ 0.15 (Ra \cdot f_2(Pr))^{1/3} & \text{for } Ra \cdot f_2(Pr) > 7 \cdot 10^4 \quad \text{(turbulent)} \end{cases} \]

\[ f_2 = \left(1 + \left(\frac{0.322}{Pr}\right)^{11/20}\right)^{20/11} \]

\[ \alpha = \frac{Nu(Gr, Pr)\lambda}{l} \]

\[ R_{th} = \frac{\Delta T}{P} = \frac{1}{\alpha A} \]

Resistor \( R_{th} \) is now calculated for one operating point! (\( \rightarrow \) more iterations are necessary!)
Example
Thermal Modeling (2)

(2) Core to Winding

\[ R_{CF} = \frac{I}{A\lambda} = 1.05 \frac{K}{W} \]

\[ R_W \ll R_{WA} \]

Heat flow via mechanical attachment has not been considered.

Measurement Results
\((\Delta B=0.18 \, T, f = 10kHz, \text{triangular})\)

<table>
<thead>
<tr>
<th></th>
<th>Calc.</th>
<th>Meas.</th>
</tr>
</thead>
<tbody>
<tr>
<td>(P_{\text{core}})</td>
<td>107 °C</td>
<td>112 °C</td>
</tr>
<tr>
<td>(P_{\text{winding}})</td>
<td>100 °C</td>
<td>104 °C</td>
</tr>
</tbody>
</table>

Quantity Values
\(R_{\text{th,CA}} = 3.6 \, K/W\)
\(R_{\text{th,WA}} = 4.7 \, K/W\)
Outline

- Magnetic Circuit Modeling
- Core Loss Modeling
- Winding Loss Modeling
- Thermal Modeling
- Multi-Objective Optimization
- Summary & Conclusion
Multi-Objective Optimization – Volume vs. Losses
Introduction to the PFC Rectifier (1)

Simplified Schematic

Converter Specifications

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input Voltage AC</td>
<td>$V_{\text{mains}}$</td>
<td>230 V</td>
</tr>
<tr>
<td>Mains Frequency</td>
<td>$f_{\text{mains}}$</td>
<td>50 Hz</td>
</tr>
<tr>
<td>DC-Voltage</td>
<td>$V_{\text{DC}}$</td>
<td>650 V</td>
</tr>
<tr>
<td>Load Current (nominal)</td>
<td>$I_L$</td>
<td>15.4 A</td>
</tr>
<tr>
<td>Switching Frequency</td>
<td>$f_{\text{sw}}$</td>
<td>8 kHz</td>
</tr>
</tbody>
</table>
Multi-Objective Optimization – Volume vs. Losses
Introduction to the PFC Rectifier (2)

**Simplified Schematic**

**Filter Specifications**

- Input Current THD ≤ 4 %
- Max. current ripple in boost inductors 4 A

**LCL filter consists of**

- three boost inductors
- and a damped three-phase LC filter.
Multi-Objective Optimization – Volume vs. Losses
Introduction to the PFC Rectifier (3)

Selected Inductor Shape

Photo of Inductor

Degrees of Freedom

\[
\begin{pmatrix}
  l_g \\
  a \\
  N \\
  do \\
  b \\
  t \\
  ww \\
  d
\end{pmatrix}
\]

Material

Grain-oriented steel
(M165-35S, lam. thickness 0.35 mm)
Multi-Objective Optimization – Volume vs. Losses
Modeling LCL Filter (1)

Procedure

1) A reluctance model is introduced to describe the electric / magnetic interface, i.e. $L = f(i)$.

2) Core losses are calculated.

3) Winding losses are calculated.

4) Inductor temperature is calculated.

Considered effects

- Air gap stray field
- Non-linearity of core material
- DC Bias
- Different flux waveforms
- Wide range of flux densities and frequencies
- Skin and proximity effect
- Stray field proximity effect
- Effect of core to magnetic field distribution
Multi-Objective Optimization – Volume vs. Losses
Modeling LCL Filter (2)

EPCOS X2 MKP Film Capacitor
(Rated voltage 305 V)

Volume Calculation

\[ 0.18 \, \mu F/cm^3 \]

Power Loss Calculation

\[ P = 2\pi f C \tan\delta V^2 \]
Multi-Objective Optimization – Volume vs. Losses
Modeling LCL Filter (3)

Simplified Schematic

Trade-off between damping capacitor size and damping achieved

\[ C = C_d \]

Optimal damping achieved with [17]

\[ R_d = \sqrt{2.1 \frac{L_1}{C}} \]

Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (1)

Simplified Schematic

Constraints concerning boost inductors

- max. current ripple $I_{HF,pp,max}$
- max. temperature $T_{max}$
- max. volume $V_{max}$
Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (2)

Simplified Schematic

Constraints concerning LC filter

max. THD of mains current
max. temperature $T_{\text{max}}$
max. volume $V_{\text{max}}$
Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (3)

Selected Inductor Shape

Filter Design Parameterization

\[
X = \begin{pmatrix}
I_{g,L1} & I_{g,L2} \\
a_{L1} & a_{L2} \\
N_{L1} & N_{L2} \\
do_{L1} & do_{L2} \\
b_{L1} & b_{L2} \\
t_{L1} & t_{L2} \\
ww_{L1} & ww_{L2} \\
d_{L1} & d_{L2}
\end{pmatrix}
\]

Filter C Calculation

\[
C = \frac{1}{L_1 \omega_0^2} = \frac{1}{L_1 \left(2\pi f_{sw} \cdot 10^{-3} \frac{A}{Hz}\right)^2}
\]

→ The filter capacitance is calculated to meet the THD constraint.
Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (4)

Simplified current / voltage waveforms for optimization procedure

Expectations

Loss overestimation in $L_2$ expected.

Loss underestimation in $L_1$ expected.
Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (5)

Optimization Flow Chart (1)

Optimization Constraints and Conditions
- max. $I_{HF,pp,max}$ in boost inductors $L_2$ 4 A
- max. THD of mains current 4 %
- max. temperature $T_{max}$ 125 °C
- max. volume $V_{max}$ 10 dm³
- switching frequency $f_{sw}$ 8 kHz
- DC link voltage $V_{DC}$ 650 V
- load current $I_L$ 15.4 A

→ Calculate $L_2$

Boost Inductor

$L_{2,min} = \frac{\sqrt{2}|V_{mains}|}{V_{DC}/\sqrt{3}} \cos(\pi/6) \cdot \frac{3}{2}V_{DC} - \sqrt{2}V_{mains} \cdot I_{HF,pp,max} \cdot f_{sw}$

$L_{2,min}$ can be calculated based on the constraint $I_{HF,pp,max}$. The maximum current ripple $I_{HF,pp,max}$ occurs when the fundamental current peaks.
Multi-Objective Optimization – Volume vs. Losses
Optimization of LCL Filter (6)

Optimization Flow Chart (2)

Cost Function

\[ F = k_{\text{Loss}}P + k_{\text{Volume}}V \]

Filter C Calculation

\[ C = \frac{1}{L_1\omega_0^2} = \frac{1}{L_1(2\pi f_{\text{sw}} \cdot 10^{9} \text{AB})^2} \]
Multi-Objective Optimization – Volume vs. Losses
Results – LCL Filter (1)

Filter Losses vs. Filter Volume
Pareto Front

\[ P \text{ [W]} \]

\[ f_{SW} = 8 \text{ kHz} \]

\[ f_{SW} = 4 \text{ kHz} \]

\[ V \text{ [dm}^3\text{]} \]
Multi-Objective Optimization – Volume vs. Losses
Results – LCL Filter (2)

Results $L_1$

**Simulated Current Waveform**

\[
THD = 3.23 \%
\]

\[
P_{\text{Loss}} = 3.0 \text{ W}
\]

**Measured Current Waveform**

\[
THD = 3.86 \%
\]

\[
P_{\text{Loss}} = 3.4 \text{ W}
\]

**Photo**

$L_1 = 230 \mu\text{H}$

Dimensions (cf. Fig. 1)

- $N = 44$
- $l_g = 4.7 \text{ mm}$
- $d_o = 60 \text{ mm}$
- $b = 60 \text{ mm}$
- $a = 20 \text{ mm}$
- $t = 8.2 \text{ mm}$
- $w_w = 56.9 \text{ mm}$
- $d = 3.15 \text{ mm}$

**Conclusion**

Loss modeling accurate.

THD underestimated
(frequency modeling necessary).
Multi-Objective Optimization – Volume vs. Losses
Results – LCL Filter (3)

Results $L_2$

- **Simulated Current Waveform**
  
  $I_{HF,pp,\text{max}} = 4.1$ A  
  $P_{\text{Loss}} = 44.9$ W

- **Measured Current Waveform**
  
  $I_{HF,pp,\text{max}} = 4.7$ A  
  $P_{\text{Loss}} = 46.8$ W

**Photo**

- Dimensions:
  - $l_1$: 1.95 mm
  - $d_0$: 60 mm
  - $b$: 60 mm
  - $a$: 20 mm
  - $t$: 28 mm
  - $w$: 56.9 mm
  - $d$: 2.24 mm

**Conclusion**

Loss modeling very accurate.

Current ripple underestimated (frequency modeling necessary).
Multi-Objective Optimization – Volume vs. Losses
Results – LCL Filter (4)

Photo of Capacitors

\[ C = C_d = 5.4 \text{ } \mu\text{F} \]
\[ R_d = 9 \text{ } \Omega \]
Multi-Objective Optimization – Volume vs. Losses
Overall System Optimization (1) – Converter Model

Converter Model

- Transistor chip area
- Diode chip area
- Cooling system volume $V_{CS}$
- Fundamental current $L_2$
- Switching frequency

Transistor junction temperature
Diode junction temperature
Converter volume (incl. DC link cap.)
Converter losses

$A_{T,D} = f(I_N)$
$P_{cond} = f(A_{T,D}, i)$
$P_{sw} \propto f_{sw}$
(extracted from data sheet)

Cooling System Performance Index (CSPI) [18]

\[
R_{th} = \frac{1}{CSPI \cdot V_{CS}}
\]

CSPI $\approx 15 \text{ W} / (\text{K liter})$

### Multi-Objective Optimization – Volume vs. Losses

**Overall System Optimization (2) – Optimization Constraints**

### Optimization Constraints

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max. junction temp.</td>
<td>$T_{j,\text{max}}$</td>
<td>125 °C</td>
</tr>
<tr>
<td>Max. cooling system vol.</td>
<td>$V_{CS,\text{max}}$</td>
<td>0.8 dm$^3$</td>
</tr>
<tr>
<td>Heatsink height</td>
<td></td>
<td>4 cm</td>
</tr>
<tr>
<td>Max. area per chip</td>
<td>$A_{T,\text{max}}/A_{D,\text{max}}$</td>
<td>1 cm$^2$</td>
</tr>
<tr>
<td>DC link voltage</td>
<td>$V_{DC}$</td>
<td>650 V</td>
</tr>
<tr>
<td>max. DC link voltage overshoot</td>
<td>$\Delta V_{DC}$</td>
<td>50 V</td>
</tr>
<tr>
<td>Fund. peak-peak current</td>
<td>$I_{(1),\text{pp}}$</td>
<td>20.5 A</td>
</tr>
</tbody>
</table>
Multi-Objective Optimization – Volume vs. Losses
Overall System Optimization (3) – Converter Pareto Front

Converter Losses vs. Converter Volume

Pareto Front

A decreasing switching frequency leads to lower losses!

Converter volume (incl. DC link cap.)
Multi-Objective Optimization – Volume vs. Losses
Overall System Optimization (3) – Optimal Designs

PFC Rectifier with Input LCL filter

Losses of loss-optimized designs
Volume of volumetric-optimized designs

→ In order to get an optimal system design, an overall system optimization has to be performed.
→ It is (often) not enough to optimize subsystems independently of each other.
Outline

- Magnetic Circuit Modeling
- Core Loss Modeling
- Winding Loss Modeling
- Thermal Modeling
- Multi-Objective Optimization

- Summary & Conclusion
Summary & Conclusion
Magnetic Circuit Modeling

Air Gap Reluctance Calculation

zy-plane

\[ R'_{zy} \]

\[ \sigma_y = \frac{R'_{zy}}{l_g} \frac{1}{\mu_0 a} \]

zx-plane

\[ R'_{zx} \]

\[ \sigma_x = \frac{R'_{zx}}{l_g} \frac{1}{\mu_0 t} \]

Results

\[ R_g = \sigma_x \sigma_y \frac{l_g}{\mu_0 a t} \]
Summary & Conclusion
Core Loss Modeling

“The best of both worlds” (Steinmetz & Loss Map approach)

\[ P_v = \frac{1}{T} \int_0^T k_i \frac{dB}{dt}^{\alpha} (\Delta B)^{\beta - \alpha} dt + \sum_{l=1}^{n} Q_{tl} P_{tl} \]

\[ \frac{P}{V} \]
Summary & Conclusion

Winding Loss Modeling

Optimal Solid Wire Thickness

Losses in Litz Wires

Foil vs. Round Conductors

Gapped cores: 2D approach
Summary & Conclusion
Magnetic Design Environment

Core Material Database

Automated Measurement System

Magnetics Design Software

Prototype

Circuit Simulator

Verification (e.g. with Calorimeter)
Summary & Conclusion
Multi-Objective Optimization - Next Steps

PFC Rectifier with Input LCL filter

Losses of loss-optimized designs

Filter Pareto Front

Next Steps

Comparison of different rectifier topologies (2-level, 3-level), modulation schemes, etc. on filter size, filter losses.
Thank you !
Additional References


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