Expert Design and Empirical Test Strategies for Practical Transformer Development

Victor W. Quinn  
Director of Engineering / CTO  
RAF Tabtronics LLC
Outline

• Units
• Design Considerations
• Coil Loss
• Magnetic Material Fundamentals
• Optimal Turns
• Empirical Shorted Load Test
• Core Loss Estimation and Recommended Empirical Test Method (iGSE,FHD,SED)
• Example Application
• Summary
New Product Development Premise
Component Verification ≠ Application Validation

- Transformer developers strive to design, develop and consistently manufacture transformers based on in depth understanding of application conditions and implementation of best practice design, production and test methods.
- The initial first article transformer must be carefully validated in the respective application circuit to ensure the scope of the component development was sufficient for the end use.
- Based on the challenges of accurate measurement of transformers in complex circuit applications, calorimetric (temperature rise) measurements are often used to characterize transformer power dissipation levels.
- The transformer developer may work closely with the power system developer to correlate loss and thermal measurements under specific conditions.
- Once the transformer design is validated, the transformer developer must implement tight QA controls on component materials and processes to ensure consistent results.
UNITs: Rationalized MKS⁺

Electric Field → \( \frac{Volt}{Meter} \)

\[ B \rightarrow \frac{\text{Webers}}{\text{Meter}^2} \Rightarrow 1 \frac{\text{Weber}}{\text{Meter}^2} = 1 \text{Tesla} = 10^4 \text{Gauss} \]

\[ H \rightarrow \frac{\text{Ampere} - \text{Turn}}{\text{Meter}} \Rightarrow 1 \frac{\text{AT}}{\text{M}} = 1.257 \times 10^{-2} \text{Oe} \]

\[ J \rightarrow \frac{\text{Ampere}}{\text{Meter}^2} \]

\[ \sigma = \frac{1}{\rho} \rightarrow \frac{1}{\text{Ohm} - \text{Meter}} \]

\[ \mu_0 = 4\pi \times 10^{-7} \frac{\text{Henry}}{\text{Meter}} = 1.257 \times 10^{-6} \frac{H}{M} \] (permeability constant)

\[ \varepsilon_0 = 8.85 \times 10^{-12} \frac{\text{Farad}}{\text{Meter}} \] (permittivity constant)

\[ \delta = \sqrt{\frac{\rho}{\pi \mu_0 f}}, \text{where } f = \text{ frequency in Hertz} \]

\[ \delta \approx \frac{2.6}{\sqrt{f}} \text{ inch (Cu @ 20 Celsius)} \]

\[ \rho \approx 0.676 \mu \Omega - \text{inch (Cu @ 20 Celsius)} \]

\[ 1 \text{ Oe} = 2.02 \text{ AT/inch} \]

\[ \mu_0 \approx 3.19 \times 10^{-8} \frac{H}{\text{inch}} \approx 0.5 \frac{G}{\text{inch}} \]

\[ \varepsilon_0 \approx 0.225 \times 10^{-12} \frac{F}{\text{inch}} \]
Design Considerations
Cross Sections and Operating Densities

Primary Section:
Coil cross section \([A_{\text{window}}]\)
constrained by aperture of core.

Secondary Section:
Core cross section \([A_{\text{core}}]\)
constrained by aperture of coil.
**Faraday’s Law**

- When a voltage $E$ is applied to a winding of $N$ Turns

$$E = (-)N \frac{d\Phi}{dt}$$

$$\Delta \Phi = \frac{1}{2N_p} \int_{t_1}^{t_1+T} |E_p| dt$$

$$|E| = \frac{1}{T} \int_{t_1}^{t_1+T} |E| dt$$

(Instantaneous Volt/Turn Equivalence)

(Average Absolute Voltage)
**Flux Density Considerations**

- Voltage driven coil

\[ \Delta \Phi = \frac{\int_{t_1}^{t_1+T} |E_p| dt}{2N_p} \]

- Line filter inductor

(Two induced consecutive, additive voltage pulses)
**Ampere’s Law**

- When a current $i$ is applied to a winding of $N$ Turns

\[ \int \vec{H} \cdot d\vec{l} = i \text{ enclosed} \]

**Curve C**

- For ideal core having infinite permeability, $H_c = 0$

\[ N_p I_p = N_s I_s \]

(Instantaneous Amp-Turn Equivalence)

*The Amp-Turn imbalance is precisely the excitation required to achieve core flux density. Magnetic material reduces the excitation current for a given flux density.*
**Transformer VA Rating**

- Faraday’s Law and Ampere’s Law imply that instantaneous Volt/Turn and Amp-Turn are constant.

- Therefore the winding Volt-Ampere product is independent of turns and is a useful parameter to rate transformers.

- Since core flux excursion is based upon average absolute winding voltage $|E|$ and winding dissipation is based upon RMS current $i_{RMS}$, define:

$$\text{Transformer } VA_{\text{Rating}} = \sum_i \left| E_i \right| \cdot i_{RMS_i}$$
**Fundamental Transformer Equation**

- Derive a relationship between Volt-Ampere rating and transformer parameters:

\[
V A_{\text{Rating}} \propto A_{\text{core}} \cdot A_{\text{window}} \cdot \text{freq} \cdot J \cdot \Delta B \cdot K_{Fe} \cdot K_{Cu}
\]

\[
V A_P \equiv A_{\text{core}} \cdot A_{\text{window}} \propto \frac{V A_{\text{Rating}}}{\text{freq} \cdot J \cdot \Delta B \cdot K_{Fe} \cdot K_{Cu}}
\]

- Therefore, by increasing frequency, current density, flux density, or utilization factors, the area product can be reduced and the transformer can be made smaller.
Current Density ($J$), Window Utilization Factor ($K_{Cu}$) and Coil Loss
Coil Loss

- At low frequencies, for uniform conductor resistivity and uniform mean lengths of turn, minimum winding loss is achieved when selected conductors yield uniform current density throughout the coil.

- For uniform current density $J$:

  \[
  \text{Low Frequency Coil Loss} = \rho \cdot \text{Cond Vol} \cdot J^2
  \]

  or

  \[
  \text{Low Frequency Coil Loss} = \Omega \cdot \left[ \sum_i N I_i \right]^2
  \]

  where

  \[
  \Omega = \frac{\rho \cdot \Delta y}{\Delta x \cdot \text{Thickness}_{cu}} = \frac{\rho \cdot \Delta y}{K_{cu} \cdot \text{Core Window Area}}
  \]

  Window Utilization Factor
Window Utilization Factor ($K_{Cu}$)

Coil Insulation Penalty

- Conductor Coatings
- Insulation Between Windings
- Insulation Between Layers
- Margins
- Core Coatings or Bobbin / Tube Supports

These considerations can total more than half of the available core window limiting utilization to less than 35%!
**Example of $K_{Cu}$ Estimate (PQ 20/20)**

\[
K_{Cu_x} = \frac{[\Delta x - 2(\text{tol}) - 2(\text{margin or bobbin flange})]}{\Delta x} = \frac{[0.55 - 2(0.003) - 2(0.036)]}{0.55} = 0.858
\]

$K_{Cu_{\text{shape}}} (\text{round versus square}) = \frac{\pi}{4} = 0.785$

$K_{Cu_{\text{insul}}} (\text{heavy film solid wire}) = 0.85$

\[
K_{Cu_z} = \frac{[\Delta z - 2(\text{tol}) - (\text{support}) - 2(2IW)]}{\Delta z} = \frac{[0.169 - 2(0.010) - (0.035) - 2(0.01)]}{0.169} = 0.556
\]

So,

$K_{Cu} \approx K_{Cu_x} \cdot K_{Cu_{\text{shape}}} \cdot K_{Cu_{\text{insul}}} \cdot K_{Cu_z} = 0.318$
**Window Utilization Penalties**

Insulation occupies more than 50% of core window.

Eddy current losses can further reduce effective conductor area by more than 50%.
Other Design Considerations Affecting Window Utilization: $K_{Cu}$
Low Frequency Example Calculation of Loss

\[ H(z)\hat{x} \equiv RH_{0\text{RMS}}\hat{x} \]

\[ \Delta z \]

\[ \hat{x} \]

\[ \hat{y} \]

\[ \hat{z} \]

\[ \text{Ampere's Law } \Rightarrow (R-1)H_{0\text{RMS}}\Delta x = i_{\text{RMS}} \]

\[ R = \frac{i_{\text{RMS}}}{H_{0\text{RMS}}\Delta x} + 1 \]

\[ J_{y\text{RMS}} = \frac{dH}{dz} = \frac{\Delta H}{\Delta z} = \frac{(R-1)H_{0\text{RMS}}}{\Delta z} = \frac{i_{\text{RMS}}}{\Delta x\Delta z} \]

\[ \text{Loss} = \rho\Delta x\Delta z \int_{y} J_{y\text{RMS}}^2 dy = \frac{\rho \Delta y i_{\text{RMS}}^2}{\Delta x\Delta z} \]

Effective conductor thickness
Qualitative Example

Severe Skin / Proximity Effects

\[
\begin{align*}
J_0 + \Delta J &= (H_0 + \Delta H) \cos(\omega t) \hat{x} \\
\Delta x &>> \Delta z \\
J_0 &= \frac{\sqrt{2}}{\delta} H_0, \quad \Delta J = \frac{\sqrt{2}}{\delta} \Delta H \\
\int_{\text{curve}1} \vec{H} \cdot d\vec{l} &= i \Rightarrow \Delta H \cos(\omega t) = \frac{i}{l_0}
\end{align*}
\]
Dissipation And Energy As Function Of B
(single layer) \( H_{s1} = 0, \ H_{s2} = H_s \)

Single layer with one surface field at zero is forgiving of design errors. However, other constructions can yield lower loss.

Minimum Dissipation @ \( B = \pi/2 \)
Loss Equivalent and Energy Equivalent Thicknesses (Single Layer Portion)

\[ \text{Power} = \frac{\rho \Delta y}{\Delta x} (\text{NI}_{\text{RMS}})^2 \cdot \left[ \frac{e^{2B} - e^{-2B} + 2 \sin(2B)}{e^{2B} + e^{-2B} - 2 \cos(2B)} \right] \frac{\Delta z}{\delta} = \frac{\rho \Delta y}{\Delta x} (\text{NI}_{\text{RMS}})^2 / \text{Thk}_{\text{equiv,loss}} \]

\[ \text{Energy} = \frac{\mu_0 \Delta y}{6 \Delta x} (\text{NI}_{\text{RMS}})^2 \cdot \frac{3}{2} \left[ \frac{e^{2B} - e^{-2B} - 2 \sin(2B)}{e^{2B} + e^{-2B} - 2 \cos(2B)} \right] \frac{\Delta z}{\delta} = \frac{\mu_0 \Delta y}{6 \Delta x} (\text{NI}_{\text{RMS}})^2 \cdot \text{Thk}_{\text{equiv,magnetic}} \]
Define Normalized Densities for General Sine Wave Boundary Conditions

\[
\text{Power} = \frac{\Delta x \Delta y \rho}{\delta} H_{0\text{RMS}}^2 p_0(R, B, \Phi) \\
\text{Magnetic Energy} = \Delta x \Delta y \mu_0 \delta H_{0\text{RMS}}^2 e_0(R, B, \Phi)
\]

- \(\Delta x \Delta y\) represents the surface area of the conductor layer
- \(H_{0\text{RMS}}\) = RMS value of sinusoidal magnetic field intensity at one boundary
- \(R\) = Ratio of magnetic surface field Intensities
- \(B\) = Ratio of conductor thickness to \(\delta\) (conductor skin depth)
- \(\Phi\) = Shift of magnetic surface field phases

 ordinless densities \((p_0, e_0)\)
Complex Winding Currents
Dissipation For Frequency Dependent Resistor

\[ V = IR \]

\[ P = I^2 R \]

\[ P_{\text{ave}} = \frac{1}{T} \int_{T} I^2 R \, dt \]

\[ P_{\text{ave}} = R \frac{1}{T} \int_{T} I^2 \, dt \]

\[ P_{\text{ave}} = R \times \text{Irms}^2 \]

\[ \text{Irms} = \sqrt{\frac{1}{T} \int_{T} I^2 \, dt} \]

For complex waveshape and frequency dependent resistor, **Harmonic Orthogonality** =>

\[ P_{\text{ave \ total}} = \sum_{n} R(f_n) \times \text{Irms}_n^2 \]
**Nonsinusoidal Excitation**

\[
\begin{align*}
I_{dc} &= \alpha_0 I_{rms} \\
I_{rms_1} &= \alpha_1 I_{rms} \\
\vdots \\
I_{rms_k} &= \alpha_k I_{rms}
\end{align*}
\]

- \( I_{rms_k} \) is the RMS current value at the kth harmonic
- \( I_{rms} \) is the RMS value of the complex waveshape
- \( \alpha_k \) is the ratio of the RMS value of the kth harmonic to the RMS value of the complex waveshape
Nonsinusoidal Currents and Multi-Layer Winding Portion

Derive equivalent power density function

\[ p_{0\text{equiv}}(R, B_1, 0) = \alpha_0^2 \frac{(1 - R)^2}{B_1} + \sum_i \sqrt{i} \alpha_i^2 p_0(R, \sqrt{i} \cdot B_1, 0) \]

Where \( B_1 \) is the ratio of conductor thickness to skin depth for the first harmonic

Sum results to derive equivalent winding thickness and effective window utilization

\[ \text{Loss Equivalent Winding Portion Thickness} = \frac{\delta}{\sum_{n=1}^{n_l} \left( \frac{n}{n_l} \right)^2 p_{0\text{equiv}} \left( 1 - \frac{1}{n}, B_1, 0 \right)} \]

\[ K_{Cu} = \frac{1}{n_l B_1 \sum_{n=1}^{n_l} \left( \frac{n}{n_l} \right)^2 p_{0\text{equiv}} \left( 1 - \frac{1}{n}, B_1, 0 \right)} \]
**Portion Utilization**

**Sine Wave, No Insulation**

For portions less than 3 skin depths thick, utilization increases with increasing layers.
**Portion Utilization**

**Sine Wave, Insulation Penalty**

Maximum conductor utilization is achieved with single layer portions, but a large number of portions is required.
Loss Equivalent Thickness
Sine Wave, No Insulation

Loss Equivalent Portion Thickness

IL = 0.0 \delta
IW = 0.0 \delta

Portion Thickness (in skin depths)

Equivalent Thickness (in skin depths)

Layers

Amplitude (normalized)

Phase (radian)

Sum of 10 Harmonics
**Fourier Decomposition Of Theoretical Waveshapes**

\[ f(t) = \sum_{n} \left[ a_n \cos(n\omega t) + b_n \sin(n\omega t) \right] \]

\[
\begin{align*}
a_n &= 0 \\
b_n &= \frac{8Ip}{DC\pi^2 n^2} \sin\left(\frac{n\pi}{2}DC\right) \\
n &= 1, 3, 5, ... \\
\end{align*}
\]

\[
\begin{align*}
a_n &= \frac{4Ip}{\pi n} \sin\left(\frac{n\pi}{2}DC\right) \\
b_n &= 0 \\
n &= 1, 3, 5, ... \\
\end{align*}
\]
Fourier Decomposition Of Theoretical Waveshapes

\[ f(t) = \sum_{n} \left[ a_n \cos(n\omega t) + b_n \sin(n\omega t) \right] \]

\[ a_n = 0 \]
\[ b_n = \frac{2I_p}{DC(1-DC)\pi^2n^2} \sin(n\pi DC) \]

for \( n = 1, 2, 3, \ldots \)

\[ a_n = \frac{2I_p}{\pi n} \sin(n\pi DC) \]
\[ b_n = 0 \]

for \( n = 1, 2, 3, \ldots \)
Loss Equivalent Thickness
Complex Wave, No Insulation

Loss Equivalent Portion Thickness

IL=0.0 δ  IW =0.0 δ

Segmentation:
- Layers: 7
  - Layer 1
  - Layer 2
  - Layer 3
  - Layer 4
  - Layer 5
  - Layer 6
  - Layer 7

Amplitude (normalized)
Phase (radian)

Sum of 10 Harmonics

Portion Thickness (in skin depths)
Loss Equivalent Thickness
Complex Wave, No Insulation

Loss Equivalent Portion Thickness

IL = 0.0 δ  IW = 0.0 δ

Layers

Sum of 10 Harmonics

Portion Thickness (in skin depths)

Equivalent Thickness (in skin depths)

Amplitude

Phase

Layers:
1, 2, 3, 4, 5, 6, 7

Sum of 10 Harmonics

Amplitude

Phase
Core Loss Introduction
Magnetic Material

- One Oersted is the value of applied field strength which causes precisely one Gauss in free space.
- Permeability is the relative gain in magnetic induction provided by magnetic material.
- Magnetic Material consists of matter spontaneously magnetized.
  - Magnetic Material in demagnetized state is divided into smaller regions called domains.
  - Each domain spontaneously magnetized to saturation level.
  - Directions of magnetization are distributed so there is no net magnetization.
  - Crystal anisotropy and stress anisotropy create preferential orientation of spontaneous magnetization within each domain.
  - Magnetostatic energy is associated with induced monopoles on surface.
  - Exchange energy resides in domain wall as a result of nonparallel adjacent atomic spin vectors.
  - Domain structure evolves to achieve relative local minimum of total stored energy.
Magnetization Microscopic Effects

- Soft magnetic material provides increased magnetic induction from induced alignment of dipoles in domain structure.
- Magnetization process converts material to single domain state magnetized in same direction as applied field.
- Magnetization process involves domain wall motion then domain magnetization rotation.
- Low frequency magnetization characteristics for magnetic material display energy storage and dissipation effects.
- Material voids and inclusions => irreversible magnetization (Br, Hc)
Optimal Transformer Turns
Design Considerations Affecting Current Density: $J$

- Maximum J
- Customer Limit
- Electrical Requirements (regulation, efficiency)
- Maximum Winding Loss
- Maximum Temperature Rise

$\text{Coil Loss } \alpha (J^2)$

Considering insulation and eddy current penalties
Design Considerations Affecting Flux Density: $\Delta B$

Core Loss $\alpha (\Delta B^\beta)$

Where exponent $\beta$ is determined through empirical core loss testing

Saturation

- $B_{peak}$ Steady State
- $B_{peak}$ Transient
- OCL Linearity

Electrical Requirements (efficiency)

Maximum Core Loss

Maximum Temperature Rise

Maximum $\Delta B$
Selection Of Transformer Parameters:  
$J$, $\Delta B$, $K_{Cu}$, $K_{Fe}$

- For given core and winding regions, $J$ and $\Delta B$ function in a competing relationship.
  - Decreasing $\Delta B$ is equivalent to increasing turns of all windings by the same proportion.
  - Increasing turns of all windings by a given proportion generates proportionately higher current density.
- For a linear transformer, selection of turns on any one winding, uniquely determines $J$ and $\Delta B$ densities and therefore total loss for a specific application.
Selection Of Transformer Parameters

- For given conductor and core regions, the normalized expression for low frequency winding loss can be extended and combined with an empirically derived core loss result to yield:

  \[ \text{Total transformer loss} = F_{\text{coil}} \cdot \frac{N_p^2}{\beta} + F_{\text{core}} \cdot \frac{N_p^\beta}{\beta} \]

- \( F_{\text{coil}} \) and \( F_{\text{core}} \) relate exponential functions of primary turns (\( N_p \)) to coil and core losses respectively.

- \( n \) is the empirically derived exponential variation of core loss with magnetic flux density.

- Without a constraint of core saturation, minimum transformer loss is achieved with primary turns:

  \[ N_p = \beta^{\frac{\beta \cdot F_{\text{core}}}{2 \cdot F_{\text{coil}}}} \]

  implying => \[ \frac{\text{Core Loss}}{\text{Winding Loss}} = \frac{2}{\beta} \]

Empirical tests characterize core material in application conditions.
Determination Of $F_{\text{coil}}$

Low Frequency Coil Loss = $\Omega \cdot \left( \sum_i N I_i \right)^2$

$\sum_i N I_i = \left[ I_p + \frac{V A_{\text{sec}}}{|V_p|} \right] N_p$

Coil Loss = $\frac{\rho \Delta y}{K_{\text{Cu}} A_{\text{window}}} \left[ I_p + \frac{V A_{\text{sec}}}{|V_p|} \right]^2 N_p^2$

$F_{\text{coil}} = \frac{\rho \Delta y}{K_{\text{Cu}} A_{\text{window}}} \left[ I_p + \frac{V A_{\text{sec}}}{|V_p|} \right]^2$

Reflects insulation and eddy current loss penalties
**Determination Of \( F_{\text{core}} \)**

\[
B = \frac{K_0 |V_p|}{fN_p A_{\text{core}}}
\]

If \( \text{Core Loss} = Volume_{Fe} K_{\text{core}} [B]^\beta \),

then \( \text{Core Loss} = Volume_{Fe} K_{\text{core}} \left( \frac{K_0 |V_p|}{fA_{\text{core}}} \right)^\beta \frac{1}{N_p^\beta} \)

and \( F_{\text{core}} = Volume_{Fe} K_{\text{core}} \left( \frac{K_0 |V_p|}{fA_{\text{core}}} \right)^\beta \)

Reflects empirical characterization of core loss for specific operating conditions and environment
Empirical Test: Verify
Transformer Realities

Winding loss

$\Phi_{\text{leakage}}$

$\Phi$

$\Phi_{\text{leakage}}$

Core Loss and Magnetic Energy applied to excite core

Leakage magnetic flux

Technology to the global power
**Coupled Coils**

The coupled coils diagram shows two coils, labeled $L_p$ and $L_s$, with currents $i_p(t)$ and $i_s(t)$, respectively. The voltage across $L_p$ is $v_p(t)$, and the voltage across $L_s$ is $v_s(t)$. The mutual inductance between the coils is $M$, and the coupling coefficient is $k$. The diagram includes a resistor $R_L$ and an impedance $Z_i$.

The equations governing the coupled coils are:

\[ v_p = L_p \frac{di_p}{dt} - M \frac{di_s}{dt} \]
\[ v_s = M \frac{di_p}{dt} - L_s \frac{di_s}{dt} \]

- $k$ is the coupling coefficient.
- $M$ is the mutual inductance.

The diagram illustrates the relationship between the voltages and currents in the coupled coils system.
Finding \( k \) and \( M \) Empirically

\[
k = \sqrt{1 - \frac{L_{sc}}{L_p}}
\]

where \( L_{sc} \) = primary inductance with secondary shorted

\[
M = k \sqrt{L_p L_s}
\]
Leakage Inductance

\[ Z_i(j\omega) = \frac{L_p}{L_s} \left[ \frac{R_L + j\omega L_s (1-k^2)}{1 - jQ_p} \right] \]

where \( Q_p \equiv \frac{R_L}{\omega L_s} \)

- Leakage inductance represented if \( Q_p << 1 \)
Lumped Series Equivalent Test Method

If $\omega L_s >> R_s$,

Primary Equivalent Leakage Inductance $= (1 - k^2) L_p$

Primary Equivalent Resistance $= R_p + k^2 \frac{L_p}{L_s} R_s$
Empirical Shorted Load Test

$$\sum_{n} I_n(nf')$$

AC Winding Loss \(\approx \sum_{n} \text{Measured Winding Loss}[I_n(nf')]\)

Average Energy \(\approx \sum_{n} \text{Measured Average Energy}[I_n(nf')]\)

Coil Loss \(\approx \text{AC Winding Loss} + \text{DC Winding Loss}\)
Core Loss Test Method Challenges

• Core magnetization is nonlinear process with nonlinear loss effects so Fourier Decomposition may be problematic
• Application waveshapes are difficult to generate
  – Rectangular voltage
  – Triangular current
  – Commercial impedance test equipment is based on sinusoidal testing at low amplitude
• Accurate loss measurement for complex waveshapes is challenging
Sine Voltage Loss Measurement Challenge

**Measured Loss Error from Uncompensated Phase Shift**

**vs Inductor Q**

Sine Wave

Q of 100 causes power measurement error ~158x phase error
Empirical Open Load Test

Parallel resonant capacitor facilitates accurate loss measurement.
Rectangular Voltage Loss Measurement Challenge

Rectangular Voltage Loss Measurement
Potential Time Delay

Time Delay
Loss Impact

Normalized Amplitude

Time (Period)
Rectangular Voltage Loss Measurement Challenge

0.05% period time delay causes 20% loss measurement error
Linear System Assumption Illustrates Loss Variation (Function of Duty Cycle)

Normalized Rectangular Wave Loss vs Duty Cycle calculated from fixed core loss resistance ($\alpha=\beta=2$)

- Fixed ET Increase
- Reverse Voltage Increase
- Duty Cycle Increase
- Nonlinear loss variation despite linear system

Graph showing normalized loss variation with duty cycle.
Modified Core Loss Estimation
Notions for Modified Core Loss Calculation Method

• Characterize a given piecewise flux excursion using a sinusoidal equivalent derivative which provides same peak flux and average slope over the interval

• The resultant induced loss by this sinusoidal equivalent derivative will be weighted by the effective duty cycle of the induced loss

• Leverage benefits of sinusoidal empirical test methods

• Facilitate consistent and application relevant transformer core loss test results
  – Correct peak flux density
  – Consistent piecewise flux derivative
Further Loss Weighting

If instantaneous loss is presumed to vary with the power of the flux derivative [1]:

\[ P(t) \propto \left| \frac{dB}{dt} \right|^\alpha \]

then a further weighting factor can be determined to scale the measured sinusoidal loss.

Note that for an exponent greater than 1, sine wave voltage generated loss is greater than square wave voltage generated loss as demonstrated historically.
**Compare Three Core Loss Estimation Methods**

- **iGSE**
  - Improved Generalized Steinmetz Equation [1]
- **FHD**
  - Fourier Harmonic Decomposition
- **SED**
  - Sinusoidal Equivalent Derivative

Initial comparisons will be based on Steinmetz power law representation as established for specific domains of sinusoidal excitation.
Comparison of Core Loss Calculation Methods

Triangular Flux Shape

Fixed Equivalent Core Loss Resistance

\[ \alpha = \beta = 2 \]

Methods consistently calculate loss for linear system
Comparison of Core Loss Calculation Methods

Triangular Flux Shape

Eddy Current Shielded Lamination

$\alpha = 1.5, \beta = 2$

Calculated Core Loss: Triangle Wave
(normalized to peak flux equivalent sine wave at fundamental frequency)

$(\alpha = 1.5, \beta = 2)$

Calculated Core Loss: Quasi Triangle Wave
(normalized to peak flux equivalent sine wave at fundamental frequency)

$(\alpha = 1.5, \beta = 2)$

Derivative Based Methods Underestimate Flux Crowding Impacts
Comparison of Core Loss Calculation Methods

Triangular Flux Shape

0.001 Permalloy 80

$\alpha = 1.71$, $\beta = 1.96$

Potential Flux Crowding Impact
Comparison of Core Loss Calculation Methods

Triangular Flux Shape

3C85 Ferrite

$\alpha = 1.5, \, \beta = 2.6$

FHD underestimates ferrite triangular loss.

Calculated Core Loss: Triangle Wave
(normalized to peak flux equivalent sine wave at fundamental frequency)

($\alpha = 1.5, \beta = 2.6$)

Calculated Core Loss: Quasi Triangle Wave
(normalized to peak flux equivalent sine wave at fundamental frequency)

($\alpha = 1.5, \beta = 2.6$)
**Harmonic Decomposition Problem**

$\alpha = 1.5$, $\beta = 2.6$

Calculated Normalized Harmonic Loss: FHD

Presumed Significant Waveshaping with Minimal Incremental Loss
Observations of Loss Methods

- iGSE and SED deliver consistent loss estimates from Steinmetz parameters since both methods presume instantaneous loss to be proportional to the power of the flux derivative.
- SED facilitates meaningful empirical test implementation beyond mathematical application of Steinmetz equation:
  - Evaluates core capacity for application driven flux density.
  - Generates application rated average winding voltages.
  - Generates rated instantaneous loss over flux excursion interval.
  - Facilitates resonant test methods for accurate and consistent results in production environment.
Evaluate Loss using Harmonic Coil
Loss Analysis and SED Core Loss Estimating Method
Example Of Coil Harmonic Loss Testing

168 W Forward Transformer 300 kHz
3 Watt Maximum Winding Dissipation
Losses By Harmonic: 16 V in

Losses By Layer

1234567
Layer Number

ACDC
Loss (Watt)

0.00.10.20.30.40.50.6

0.1

Tabtronics: 568-6078, 16 V in,
53% Duty Cycle

Calculated
Actual (5 harmonics)
Loss Summary
168 W Forward Converter

- DC Coil: 0.5 W
- AC Coil: 1 W
- AC Core: 1.5 W
- Total: 3 W

8 V, 21 A, 300 kHz
Other Observed Core Loss Factors

- Fundamental variables (peak field density, frequency, temperature)
- Biased operating application waveshape [2]
- Flux density relaxation during circuit commutation intervals
- Flux crowding from eddy current shielding
- Designed air gap => fringing effects
- Nonuniform core cross sections => nonuniform magnetic field densities => nonuniform loss density
- Unintended micro-crack(s)
- Mechanical Stress (magnetic anisotropy)
- Irregular or mismatched core segment interfaces
- Material manufacturing lot variances (unintended substitutions !)
Core Loss Experiences

Transformer Loss (PSFB, 214W, 100 kHz)

Table 1: expected performance comparison on ungapped core set

<table>
<thead>
<tr>
<th>Material</th>
<th>NLP Terminals (4-5) 350Vrms @ 200kHz</th>
<th>NLP Terminals (4-5) 350Vrms @ 200kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>3C94</td>
<td>20°C: 1.92W</td>
<td>100°C: 1.022W</td>
</tr>
<tr>
<td>3C96</td>
<td>20°C: 2.05W</td>
<td>100°C: 0.74W</td>
</tr>
<tr>
<td>(3C96 - 3C94)</td>
<td>20°C: 0.13W</td>
<td>100°C: -0.282W</td>
</tr>
</tbody>
</table>

Micro-cracks – Misalignment >3x core loss increase
Technology to the global power

Summary
Recommendations

• **Rigorously measure performance of magnetic components used in demanding applications prior to commencing production.**
  – **Use application relevant high frequency and high amplitude test methods to ensure consistent application results throughout production.**

• If the magnetics have significant impact on the end system’s market differentiation then a collaborative *Design Partnership* between power system developer and transformer developer will maximize the custom magnetics’ utility for the end system.
  – **Capable and trustworthy design partners with effective project management is necessary.**
References


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