14.3 Multiple-winding magnetics design using the $K_g$ method

The $K_g$ design method can be extended to multiple-winding magnetic elements such as transformers and coupled inductors.

This method is applicable when

- Copper loss dominates the total loss (i.e. core loss is ignored), or
- The maximum flux density $B_{\text{max}}$ is a specification rather than a quantity to be optimized

To do this, we must

- Find how to allocate the window area between the windings
- Generalize the step-by-step design procedure
14.3.1 Window area allocation

**Given:** application with $k$ windings having known rms currents and desired turns ratios

\[ \frac{v_1(t)}{n_1} = \frac{v_2(t)}{n_2} = \ldots = \frac{v_k(t)}{n_k} \]

Given:
- Application with $k$ windings having known rms currents and desired turns ratios
- \[ \frac{v_1(t)}{n_1} = \frac{v_2(t)}{n_2} = \ldots = \frac{v_k(t)}{n_k} \]

**Q:** how should the window area $W_A$ be allocated among the windings?
Allocation of winding area

\[
\begin{align*}
\text{Winding 1 allocation} & \quad \alpha_1 W_A \\
\text{Winding 2 allocation} & \quad \alpha_2 W_A \\
\text{etc.} & \\
\end{align*}
\]

\[
0 < \alpha_j < 1 \\
\alpha_1 + \alpha_2 + \cdots + \alpha_k = 1
\]
Copper loss in winding $j$

Copper loss (not accounting for proximity loss) is

$$P_{cu,j} = I_j^2 R_j$$

Resistance of winding $j$ is

$$R_j = \rho \frac{\ell_j}{A_{W,j}}$$

with

$$\ell_j = n_j (MLT)$$  \hspace{1cm} \text{length of wire, winding } j$$

$$A_{W,j} = \frac{W_A K_u \alpha_j}{n_j}$$  \hspace{1cm} \text{wire area, winding } j$$

Hence

$$R_j = \rho \frac{n_j^2 (MLT)}{W_A K_u \alpha_j}$$

$$P_{cu,j} = \frac{n_j^2 i_j^2 \rho (MLT)}{W_A K_u \alpha_j}$$
Total copper loss of transformer

Sum previous expression over all windings:

\[ P_{cu,tot} = P_{cu,1} + P_{cu,2} + \cdots + P_{cu,k} = \rho \frac{(MLT)}{W_A K_u} \sum_{j=1}^{k} \left( \frac{n_j^2 I_j^2}{\alpha_j} \right) \]

Need to select values for \( \alpha_1, \alpha_2, \ldots, \alpha_k \) such that the total copper loss is minimized
Variation of copper losses with $\alpha_1$

For $\alpha_1 = 0$: wire of winding 1 has zero area. $P_{cu,1}$ tends to infinity

For $\alpha_1 = 1$: wires of remaining windings have zero area. Their copper losses tend to infinity

There is a choice of $\alpha_1$ that minimizes the total copper loss
Method of Lagrange multipliers
to minimize total copper loss

Minimize the function

\[ P_{cu,\text{tot}} = P_{cu,1} + P_{cu,2} + \cdots + P_{cu,k} = \frac{\rho \cdot (MLT)}{W_A K_u} \sum_{j=1}^{k} \left( \frac{n_j^2 I_j^2}{\alpha_j} \right) \]

subject to the constraint

\[ \alpha_1 + \alpha_2 + \cdots + \alpha_k = 1 \]

Define the function

\[ f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi) = P_{cu,\text{tot}}(\alpha_1, \alpha_2, \ldots, \alpha_k) + \xi \cdot g(\alpha_1, \alpha_2, \ldots, \alpha_k) \]

where

\[ g(\alpha_1, \alpha_2, \ldots, \alpha_k) = 1 - \sum_{j=1}^{k} \alpha_j \]

is the constraint that must equal zero

and \( \xi \) is the Lagrange multiplier.
Lagrange multipliers
continued

Optimum point is solution of the system of equations
\[
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_1} = 0
\]
\[
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_2} = 0
\]
\[
\vdots
\]
\[
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_k} = 0
\]
\[
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \xi} = 0
\]

Result:
\[
\xi = \frac{\rho (MLT)}{W_A K_u} \left( \sum_{j=1}^{k} n_j I_j \right)^2 = P_{cu,tot}
\]
\[
\alpha_m = \frac{n_m I_m}{\sum_{n=1}^{\infty} n_j I_j}
\]

An alternate form:
\[
\alpha_m = \frac{V_m I_m}{\sum_{n=1}^{\infty} V_j I_j}
\]
Interpretation of result

\[ \alpha_m = \frac{V_m I_m}{\sum_{n=1}^{\infty} V_j I_j} \]

Apparent power in winding \(j\) is

\[ V_j I_j \]

where \(V_j\) is the rms or peak applied voltage

\(I_j\) is the rms current

Window area should be allocated according to the apparent powers of the windings
Example
PWM full-bridge transformer

- Note that waveshapes (and hence rms values) of the primary and secondary currents are different.
- Treat as a three-winding transformer.
Expressions for RMS winding currents

\[ I_1 = \sqrt{\frac{1}{2T_s} \int_0^{2T_s} i_1^2(t) \, dt} = \frac{n_2}{n_1} I \sqrt{D} \]

\[ I_2 = I_3 = \sqrt{\frac{1}{2T_s} \int_0^{2T_s} i_2^2(t) \, dt} = \frac{1}{2} I \sqrt{1 + D} \]

see Appendix A
Allocation of window area: \[ \alpha_m = \frac{V_m I_m}{\sum_{n=1}^{\infty} V_j I_j} \]

Plug in rms current expressions. Result:

\[ \alpha_1 = \frac{1}{\left(1 + \sqrt{\frac{1 + D}{D}}\right)} \]  
\[ \alpha_2 = \alpha_3 = \frac{1}{2} \frac{1}{\left(1 + \sqrt{\frac{D}{1 + D}}\right)} \]

Fraction of window area allocated to primary winding
Fraction of window area allocated to each secondary winding
Numerical example

Suppose that we decide to optimize the transformer design at the worst-case operating point $D = 0.75$. Then we obtain

$$
\alpha_1 = 0.396 \\
\alpha_2 = 0.302 \\
\alpha_3 = 0.302
$$

The total copper loss is then given by

$$
P_{cu,\text{tot}} = \frac{\rho(MLT)}{W_A K_u} \left( \sum_{j=1}^{3} n_j I_j \right)^2 \\
= \frac{\rho(MLT)n_2^2 I^2}{W_A K_u} \left( 1 + 2D + 2\sqrt{D(1+D)} \right)
$$
14.3.2 Coupled inductor design constraints

Consider now the design of a coupled inductor having $k$ windings. We want to obtain a specified value of magnetizing inductance, with specified turns ratios and total copper loss.

Magnetic circuit model:
Relationship between magnetizing current and winding currents

Solution of circuit model, or by use of Ampere’s Law:

\[ i_M(t) = i_1(t) + \frac{n_2}{n_1} i_2(t) + \cdots + \frac{n_k}{n_1} i_k(t) \]
Solution of magnetic circuit model:
Obtain desired maximum flux density

Design so that the maximum flux density $B_{\text{max}}$ is equal to a specified value (that is less than the saturation flux density $B_{\text{sat}}$). $B_{\text{max}}$ is related to the maximum magnetizing current according to

$$n_1 I_{M,\text{max}} = B_{\text{max}} A_c R_g = B_{\text{max}} \frac{\ell_g}{\mu_0}$$
Obtain specified magnetizing inductance

By the usual methods, we can solve for the value of the magnetizing inductance $L_M$ (referred to the primary winding):

$$L_M = \frac{n_1^2}{\mathcal{R}_g} = n_1^2 \frac{\mu_0 A_c}{\ell_g}$$
Copper loss

Allocate window area as described in Section 14.3.1. As shown in that section, the total copper loss is then given by

\[ P_{cu} = \frac{\rho (MLT)n_1^2 I_{tot}^2}{W_A K_u} \]

with

\[ I_{tot} = \sum_{j=1}^{k} \frac{n_j}{n_1} I_j \]
Eliminate unknowns and solve for $K_g$

Eliminate the unknowns $\ell_g$ and $n_1$:

$$P_{cu} = \frac{\rho (MLT)L_M^2 I_{tot}^2 I_{M,max}^2}{B_{max}^2 A_c^2 W_A K_u}$$

Rearrange equation so that terms that involve core geometry are on RHS while specifications are on LHS:

$$\frac{A_c^2 W_A}{(MLT)} = \frac{\rho L_M^2 I_{tot}^2 I_{M,max}^2}{B_{max}^2 K_u P_{cu}}$$

The left-hand side is the same $K_g$ as in single-winding inductor design. Must select a core that satisfies

$$K_g \geq \frac{\rho L_M^2 I_{tot}^2 I_{M,max}^2}{B_{max}^2 K_u P_{cu}}$$
14.3.3 Step-by-step design procedure: Coupled inductor

The following quantities are specified, using the units noted:

- Wire resistivity $\rho$ (Ω-cm)
- Total rms winding currents $I_{tot} = \sum_{j=1}^{k} \frac{n_j}{n_1} I_j$ (A) (referred to winding 1)
- Peak magnetizing current $I_{M,max}$ (A) (referred to winding 1)
- Desired turns ratios $n_2/n_1, n_3/n_2, \text{etc.}$
- Magnetizing inductance $L_M$ (H) (referred to winding 1)
- Allowed copper loss $P_{cu}$ (W)
- Winding fill factor $K_u$
- Core maximum flux density $B_{max}$ (T)

The core dimensions are expressed in cm:

- Core cross-sectional area $A_c$ (cm$^2$)
- Core window area $W_A$ (cm$^2$)
- Mean length per turn $MLT$ (cm)

The use of centimeters rather than meters requires that appropriate factors be added to the design equations.
1. Determine core size

\[ K_g \geq \frac{\rho L_M^2 I_{tot}^2 I_{M,\text{max}}^2}{B_{\text{max}}^2 P_{\text{cu}} K_u} \times 10^8 \quad \text{(cm}^5) \]

Choose a core that satisfies this inequality. Note the values of \( A_c \), \( W_A \), and \( \text{MLT} \) for this core.

The resistivity \( \rho \) of copper wire is \( 1.724 \times 10^{-6} \, \Omega \, \text{cm} \) at room temperature, and \( 2.3 \times 10^{-6} \, \Omega \, \text{cm} \) at \( 100^\circ \text{C} \).
2. Determine air gap length

\[ \ell_g = \frac{\mu_0 L_M I_{M,\text{max}}^2}{B_{\text{max}}^2 A_c} \times 10^4 \text{ (m)} \]

(value neglects fringing flux, and a longer gap may be required)

The permeability of free space is \( \mu_0 = 4\pi \cdot 10^{-7} \text{ H/m} \)
3. Determine number of turns

For winding 1:

\[ n_1 = \frac{L_M I_{M,\text{max}}}{B_{\text{max}} A_c} \times 10^4 \]

For other windings, use the desired turns ratios:

\[ n_2 = \left( \frac{n_2}{n_1} \right) n_1 \]
\[ n_3 = \left( \frac{n_3}{n_1} \right) n_1 \]
\[ \vdots \]
4. Evaluate fraction of window area allocated to each winding

\[
\alpha_1 = \frac{n_1 I_1}{n_1 I_{tot}}
\]

\[
\alpha_2 = \frac{n_2 I_2}{n_1 I_{tot}}
\]

\[
\vdots
\]

\[
\alpha_k = \frac{n_k I_k}{n_1 I_{tot}}
\]

\[
\begin{align*}
\text{Winding 1 allocation} & : \alpha_1 W_A \\
\text{Winding 2 allocation} & : \alpha_2 W_A \\
\text{etc.} & \\
\end{align*}
\]

Total window area \( W_A \)

\[
0 < \alpha_j < 1 \\
\alpha_1 + \alpha_2 + \cdots + \alpha_k = 1
\]
5. Evaluate wire sizes

\[ A_{w1} \leq \frac{\alpha_1 K_u W_A}{n_1} \]

\[ A_{w2} \leq \frac{\alpha_2 K_u W_A}{n_2} \]

\[ \vdots \]

See American Wire Gauge (AWG) table at end of Appendix D.
14.4 Examples

14.4.1 Coupled Inductor for a Two-Output Forward Converter

14.4.2 CCM Flyback Transformer
14.4.1 Coupled Inductor for a Two-Output Forward Converter

The two filter inductors can share the same core because their applied voltage waveforms are proportional. Select turns ratio $n_2/n_1$ approximately equal to $v_2/v_1 = 12/28$. 

Output 1
- 28 V
- 4 A

Output 2
- 12 V
- 2 A

$f_s = 200$ kHz
Coupled inductor model and waveforms

Secondary-side circuit, with coupled inductor model

Magnetizing current and voltage waveforms. \(i_M(t)\) is the sum of the winding currents \(i_1(t) + i_2(t)\).
Nominal full-load operating point

Design for CCM operation with
\[ D = 0.35 \]
\[ \Delta i_M = 20\% \text{ of } I_M \]
\[ f_s = 200 \text{ kHz} \]

DC component of magnetizing current is

\[ I_M = I_1 + \frac{n_2}{n_1} I_2 \]
\[ = (4 \text{ A}) + \frac{12}{28} (2 \text{ A}) \]
\[ = 4.86 \text{ A} \]
Magnetizing current ripple

\[ \Delta i_M = \frac{V_1 D' T_s}{2L_M} \]

To obtain \( \Delta i_M = 20\% \) of \( I_M \)

choose

\[ L_M = \frac{V_1 D' T_s}{2\Delta i_M} \]

\[ = \frac{(28 \text{ V})(1 - 0.35)(5 \mu\text{s})}{2(4.86 \text{ A})(20\%)} \]

\[ = 47 \mu\text{H} \]

This leads to a peak magnetizing current (referred to winding 1) of

\[ I_{M,\text{max}} = I_M + \Delta i_M = 5.83 \text{ A} \]
RMS winding currents

Since the winding current ripples are small, the rms values of the winding currents are nearly equal to their dc components:

\[ I_1 = 4 \text{ A} \quad I_2 = 2 \text{ A} \]

Hence the sum of the rms winding currents, referred to the primary, is

\[ I_{tot} = I_1 + \frac{n_2}{n_1} I_2 = 4.86 \text{ A} \]
Evaluate $K_g$

The following engineering choices are made:

- Allow 0.75 W of total copper loss (a small core having thermal resistance of less than 40 °C/W then would have a temperature rise of less than 30 °C)
- Operate the core at $B_{max} = 0.25$ T (which is less than the ferrite saturation flux density of 0.3 to 0.5 T)
- Use fill factor $K_u = 0.4$ (a reasonable estimate for a low-voltage inductor with multiple windings)

Evaluate $K_g$:

$$K_g \geq \left( \frac{1.724 \cdot 10^{-6} \ \Omega \text{ cm}}{47 \ \mu\text{H}} \right)^2 \left( \frac{4.86 \ \text{A}}{5.83 \ \text{A}} \right)^2 \left( \frac{0.25 \ \text{T}}{0.75 \ \text{W}} \right)^2 \left( \frac{0.75 \ \text{W}}{0.4} \right) 10^8$$

$$= 16 \cdot 10^{-3} \ \text{cm}^5$$
Select core

It is decided to use a ferrite PQ core. From Appendix D, the smallest PQ core having $K_g \geq 16 \cdot 10^{-3} \text{ cm}^5$ is the PQ 20/16, with $K_g = 22.4 \cdot 10^{-3} \text{ cm}^5$. The data for this core are:

$$A_c = 0.62 \text{ cm}^2$$

$$W_A = 0.256 \text{ cm}^2$$

$$MLT = 4.4 \text{ cm}$$
Air gap length

\[ \ell_g = \frac{\mu_0 L_M I_{M,\text{max}}^2}{B_{\text{max}}^2 A_c} \times 10^4 \]

\[ = \frac{(4\pi \cdot 10^{-7}\text{H/m})(47 \mu\text{H})(5.83 \text{A})^2}{(0.25 \text{T})^2(0.62 \text{cm}^2)} \times 10^4 \]

\[ = 0.52 \text{ mm} \]
Turns

\[ n_1 = \frac{L_M I_{M,\text{max}}}{B_{\text{max}} A_c} \times 10^4 \]
\[ = \frac{(47 \ \mu\text{H})(5.83 \ \text{A})}{(0.25 \ \text{T})(0.62 \ \text{cm}^2)} \times 10^4 \]
\[ = 17.6 \text{ turns} \]

\[ n_2 = \left( \frac{n_2}{n_1} \right) n_1 \]
\[ = \left( \frac{12}{28} \right) (17.6) \]
\[ = 7.54 \text{ turns} \]

Let’s round off to

\[ n_1 = 17 \]
\[ n_2 = 7 \]
Wire sizes

Allocation of window area:

\[
\alpha_1 = \frac{n_1 I_1}{n_1 I_{tot}} = \frac{(17)(4 \text{ A})}{(17)(4.86 \text{ A})} = 0.8235
\]

\[
\alpha_2 = \frac{n_2 I_2}{n_1 I_{tot}} = \frac{(7)(2 \text{ A})}{(17)(4.86 \text{ A})} = 0.1695
\]

Determination of wire areas and AWG (from table at end of Appendix D):

\[
A_{w1} \leq \frac{\alpha_1 K_u W_A}{n_1} = \frac{(0.8235)(0.4)(0.256 \text{ cm}^2)}{(17)} = 4.96 \cdot 10^{-3} \text{ cm}^2
\]

use AWG #21

\[
A_{w2} \leq \frac{\alpha_2 K_u W_A}{n_2} = \frac{(0.1695)(0.4)(0.256 \text{ cm}^2)}{(7)} = 2.48 \cdot 10^{-3} \text{ cm}^2
\]

use AWG #24
14.4.2 Example 2: CCM flyback transformer

Transformer model: $n_1 : n_2$

$i_M$ and $i_2$ current waveforms

$\Delta i_M$ is the change in $i_M$

$\frac{n_1}{n_2} I_M$ is the secondary current

$V_g$ is the input voltage

$DT_s$ is the duty cycle time
## Specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>$V_g = 200\text{V}$</td>
</tr>
<tr>
<td>Output (full load)</td>
<td>20 V at 5 A</td>
</tr>
<tr>
<td>Switching frequency</td>
<td>150 kHz</td>
</tr>
<tr>
<td>Magnetizing current ripple</td>
<td>20% of dc magnetizing current</td>
</tr>
<tr>
<td>Duty cycle</td>
<td>$D = 0.4$</td>
</tr>
<tr>
<td>Turns ratio</td>
<td>$n_2/n_1 = 0.15$</td>
</tr>
<tr>
<td>Copper loss</td>
<td>1.5 W</td>
</tr>
<tr>
<td>Fill factor</td>
<td>$K_u = 0.3$</td>
</tr>
<tr>
<td>Maximum flux density</td>
<td>$B_{\text{max}} = 0.25 \text{T}$</td>
</tr>
</tbody>
</table>
Basic converter calculations

Components of magnetizing current, referred to primary:

\[ I_M = \left( \frac{n_2}{n_1} \right) \frac{1}{D'} \frac{V}{R} = 1.25 \text{ A} \]

\[ \Delta i_M = (20\%)I_M = 0.25 \text{ A} \]

\[ I_{M,max} = I_M + \Delta i_M = 1.5 \text{ A} \]

Choose magnetizing inductance:

\[ L_M = \frac{V_g DT_s}{2\Delta i_M} \]
\[ = 1.07 \text{ mH} \]

RMS winding currents:

\[ I_1 = I_M \sqrt{D} \sqrt{1 + \frac{1}{3} \left( \frac{\Delta i_M}{I_M} \right)^2} = 0.796 \text{ A} \]

\[ I_2 = \frac{n_1}{n_2} I_M \sqrt{D'} \sqrt{1 + \frac{1}{3} \left( \frac{\Delta i_M}{I_M} \right)^2} = 6.50 \text{ A} \]

\[ I_{tot} = I_1 + \frac{n_2}{n_1} I_2 = 1.77 \text{ A} \]
Choose core size

\[ K_g \geq \frac{\rho L_M^2 I_{tot}^2 I_{M,max}^2}{B_{max}^2 P_{cu} K_u} \times 10^8 \]

\[ = \frac{\left(1.724 \cdot 10^{-6} \Omega \cdot \text{cm}\right) \left(1.07 \cdot 10^{-3} \text{ H}\right)^2 \left(1.77 \text{ A}\right)^2 \left(1.5 \text{ A}\right)^2}{\left(0.25 \text{ T}\right)^2 \left(1.5 \text{ W}\right) \left(0.3\right)} \times 10^8 \]

\[ = 0.049 \text{ cm}^5 \]

The smallest EE core that satisfies this inequality (Appendix D) is the EE30.
Choose air gap and turns

\[ \ell_g = \frac{\mu_0 L_M I_{M,\text{max}}^2}{B_{\text{max}}^2 A_c} \times 10^4 \]

\[ = \frac{\left(4\pi \cdot 10^{-7} \text{H/m}\right)\left(1.07 \cdot 10^{-3} \text{ H}\right)\left(1.5 \text{ A}\right)^2}{\left(0.25 \text{ T}\right)^2\left(1.09 \text{ cm}^2\right)} \times 10^4 \]

\[ = 0.44 \text{ mm} \]

\[ n_1 = \frac{L_M I_{M,\text{max}}}{B_{\text{max}} A_c} \times 10^4 \]

\[ = \frac{\left(1.07 \cdot 10^{-3} \text{ H}\right)\left(1.5 \text{ A}\right)}{\left(0.25 \text{ T}\right)\left(1.09 \text{ cm}^2\right)} \times 10^4 \]

\[ = 58.7 \text{ turns} \]

Round to \( n_1 = 59 \)

\[ n_2 = \left(\frac{n_2}{n_1}\right) n_1 \]

\[ = \left(0.15\right) 59 \]

\[ = 8.81 \]

\[ = 9 \]
Wire gauges

\[
\alpha_1 = \frac{I_1}{I_{\text{tot}}} = \frac{(0.796 \text{ A})}{(1.77 \text{ A})} = 0.45
\]

\[
\alpha_2 = \frac{n_2 I_2}{n_1 I_{\text{tot}}} = \frac{(9)(6.5 \text{ A})}{(59)(1.77 \text{ A})} = 0.55
\]

\[
A_{W1} \leq \frac{\alpha_1 K_u W_A}{n_1} = 1.09 \cdot 10^{-3} \text{ cm}^2 \quad \text{— use #28 AWG}
\]

\[
A_{W2} \leq \frac{\alpha_2 K_u W_A}{n_2} = 8.88 \cdot 10^{-3} \text{ cm}^2 \quad \text{— use #19 AWG}
\]
Core loss
CCM flyback example

B-H loop for this application:

The relevant waveforms:

\[
\frac{dB(t)}{dt} = \frac{v_M(t)}{n_1 A_c}
\]

For the first subinterval:

\[
\frac{dB(t)}{dt} = \frac{V_g}{n_1 A_c}
\]
Calculation of ac flux density and core loss

Solve for $\Delta B$:

$$\Delta B = \left( \frac{V_g}{n_1 A_c} \right) (DT_s)$$

Plug in values for flyback example:

$$\Delta B = \frac{(200 \text{ V})(0.4)(6.67 \mu\text{s})}{2(59)(1.09 \text{ cm}^2)} \times 10^4$$

$$= 0.041 \text{ T}$$

From manufacturer’s plot of core loss (at left), the power loss density is $0.04 \text{ W/cm}^3$. Hence core loss is

$$P_{fe} = \left( 0.04 \text{ W/cm}^3 \right) \left( A_c \ell_m \right)$$

$$= \left( 0.04 \text{ W/cm}^3 \right) \left( 1.09 \text{ cm}^2 \right) \left( 5.77 \text{ cm} \right)$$

$$= 0.25 \text{ W}$$
Comparison of core and copper loss

- Copper loss is 1.5 W
  - does not include proximity losses, which could substantially increase total copper loss
- Core loss is 0.25 W
  - Core loss is small because ripple and $\Delta B$ are small
  - It is not a bad approximation to ignore core losses for ferrite in CCM filter inductors
  - Could consider use of a less expensive core material having higher core loss
  - Neglecting core loss is a reasonable approximation for this application
- Design is dominated by copper loss
  - The dominant constraint on flux density is saturation of the core, rather than core loss