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

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

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

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

Core
Window area $W_A$
Core mean length per turn (MLT)
Wire resistivity $\rho$
Fill factor $K_f$
Allocation of winding area

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

\[
0 < \alpha_j < 1 \\
\alpha_1 + \alpha_2 + \cdots + \alpha_k = 1
\]
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

$$\ell_j = n_j (MLT)$$

length of wire, winding $j$

$$A_{W,j} = \frac{W_A K_u \alpha_j}{n_j}$$

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

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

subject to the constraint

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

Define the function

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

where

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

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
\[
\begin{align*}
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_1} &= 0 \\
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_2} &= 0 \\
\vdots \\
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \alpha_k} &= 0 \\
\frac{\partial f(\alpha_1, \alpha_2, \ldots, \alpha_k, \xi)}{\partial \xi} &= 0
\end{align*}
\]

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

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

Fraction of window area allocated to primary winding

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

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

Assume that gap reluctance is much larger than core reluctance:

\[ n_1 i_M(t) = B(t) A_c R_g \]

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

\[ n_1 I_{M,\text{max}} = B_{\text{max}} A_c R_g = B_{\text{max}} \frac{L_g}{\mu_0} \]
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):

$$L_M = \frac{n_1^2}{\mathcal{R}_g} = n_1^2 \frac{\mu_0 A_c}{l_g}$$
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

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

Eliminate the unknowns $l_g$ and $n_1$:

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

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

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

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

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

The following quantities are specified, using the units noted:

- **Wire resistivity** \( \rho \) (\( \Omega \cdot \text{cm} \))
- **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** \( I_{M,\text{max}} \) (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** \( P_{\text{cu}} \) (W)
- **Winding fill factor** \( K_u \)
- **Core maximum flux density** \( B_{\text{max}} \) (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

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

Choose a core that satisfies this inequality. Note the values of \( A_c \), \( W_A \), and \( MLT \) 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°C.
2. Determine air gap length

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

(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

\[ \alpha_1 = \frac{n_1 I_1}{n_1 I_{tot}} \]
\[ \alpha_2 = \frac{n_2 I_2}{n_1 I_{tot}} \]
\[ \vdots \]
\[ \alpha_k = \frac{n_k I_k}{n_1 I_{tot}} \]

Winding 1 allocation \( \alpha_1 W_A \)

Winding 2 allocation \( \alpha_2 W_A \)

\( \text{etc.} \)

\[ 0 < \alpha_j < 1 \]
\[ \alpha_1 + \alpha_2 + \cdots + \alpha_k = 1 \]
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

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

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

\[ I_{M,\max} = I_M + \Delta i_M = 5.83 \text{ A} \]
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 $B_{\text{max}} = 0.25$ 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 $K_g$:

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

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

- $A_c = 0.62$ cm$^2$
- $W_A = 0.256$ cm$^2$
- $MLT = 4.4$ cm
Air gap length

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

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

\[ = 0.52 \text{ mm} \]
Inductor design

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

\[ = \frac{(47 \, \mu H)(5.83 \, A)}{(0.25 \, T)(0.62 \, \text{cm}^2)} \times 10^4 \]

\[ = 17.6 \text{ turns} \]

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

\[ = \left( \frac{12}{28} \right) (17.6) \]

\[ = 7.54 \text{ turns} \]

Let’s round off to

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

Allocation of window area:

\[
\begin{align*}
\alpha_1 &= \frac{n_1 I_1}{n_1 I_{tot}} = \frac{(17)(4 \text{ A})}{(17)(4.86 \text{ A})} = 0.8235 \\
\alpha_2 &= \frac{n_2 I_2}{n_1 I_{tot}} = \frac{(7)(2 \text{ A})}{(17)(4.86 \text{ A})} = 0.1695
\end{align*}
\]

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

\[ i_M(t) = 0 \]

\[ i_2(t) = \frac{n_1}{n_2} I_M \]

\[ v_M(t) = 0 \]

\[ V_g = DT_i \]

\[ Q_1 \]

\[ L_M \]

\[ V \]

\[ R \]

\[ C \]
## Specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input voltage</td>
<td>$V_g = 200$ 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>$B_{max} = 0.25$ T</td>
</tr>
</tbody>
</table>
Basic converter calculations

Components of magnetizing current, referred to primary:

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

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

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

Choose magnetizing inductance:

\[ L_M = \frac{V_e DT_s}{2 \Delta i_M} \]

\[ = 1.07 \text{ mH} \]

RMS winding currents:

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

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

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

\[
K_g \geq \frac{\rho L_M^2 I_{tot}^2 I_{M,\max}^2}{B_{\max}^2 P_{cu} K_u} 10^8
= \frac{\left(1.724 \cdot 10^{-6} \Omega \text{-cm}\right) \left(1.07 \cdot 10^{-3} \text{ H}\right)^2 \left(1.77 \text{ A}\right)^2 \left(1.5 \text{ A}\right)^2}{\left(0.25 \text{ T}\right)^2 \left(1.5 \text{ W}\right) \left(0.3\right)} 10^8
= 0.049 \text{ cm}^5
\]

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

\[ l_g = \frac{\mu_0 L_M I_{M,\text{max}}^2}{B_{\text{max}}^2 A_c} \times 10^4 \]
\[ = \frac{(4\pi \times 10^{-7} \text{ H/m}) \times (1.07 \times 10^{-3} \text{ H}) \times (1.5 \text{ A})^2}{(0.25 \text{ T})^2 \times (1.09 \text{ cm}^2)} \times 10^4 \]
\[ = 0.44 \text{ mm} \]

\[ n_1 = \frac{L_M I_{M,\text{max}}}{B_{\text{max}} A_c} \times 10^4 \]
\[ = \frac{(1.07 \times 10^{-3} \text{ H}) \times (1.5 \text{ A})}{(0.25 \text{ T}) \times (1.09 \text{ cm}^2)} \times 10^4 \]
\[ = 58.7 \text{ turns} \]

Round to \( n_1 = 59 \)

\[ n_2 = \frac{n_2}{n_1} \times n_1 \]
\[ = (0.15) \times 59 \]
\[ = 8.81 \]

\[ n_2 = 9 \]
Wire gauges

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

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

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

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

B-H loop for this application:

 ![B-H loop diagram]

The relevant waveforms:

For the first subinterval:
\[
\frac{dB(t)}{dt} = \frac{V_g}{n_1 A_c} 
\]

\( B(t) \) vs. applied voltage, from Faraday’s law:
\[
\frac{dB(t)}{dt} = \frac{v_M(t)}{n_1 A_c}
\]

Fundamentals of Power Electronics

Chapter 14: Inductor design
Calculation of ac flux density and core loss

Solve for $\Delta B$:

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

Plug in values for flyback example:

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

$$= 0.041 \text{ T}$$

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

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

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

$$= 0.25 \text{ W}$$
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