Iron Loss Calculation in a Claw-pole Structure

A. Reinap, D. Martínez-Muñoz, and M. Alaküla

Abstract—The total core loss in a claw-pole structure has been calculated on the basis of the variation of the magnetic loading over the magnetization cycle in different parts of the machine. This variation has been estimated from a lumped parameter permeance network and 3D finite element simulations, studying the flux density waveform and loci in their elements. The results from the simulations are compared with the loss measurements in a single-phase claw-pole motor equipped with an outer permanent-magnet rotor. Results show that the static and dynamic losses are underestimated by 35% and up to 60% at 100Hz respectively. Finally, the results from an optimization process are presented, where the torque performance is evaluated as a function of the pole number and dimensions of the machine.

Index Terms—3D finite element field computation, claw-pole machines, core losses, lumped parameter permeance network, optimization, soft magnetic composite (SMC) materials.

I. INTRODUCTION

THE use of soft magnetic composites (SMC) and compaction technology allows creating complex 3D ferrous structures, such as claw-poles. In these structures the core is subject to a 3D variation of the magnetic flux over an excitation cycle, and the trajectory of the flux vector forms different shapes in different parts of the core. It has been shown that these effects have to be considered in the loss model for an accurate prediction of the losses [1]-[3].

The magnetic equivalent circuit (MEC) has often been introduced as a fast design model for the calculation of the optimal size of the core [4], [5]. Even though the model of 1D elements describes roughly the magnetic flux flow in the 3D object, the approximate sizing is enough to initialize the geometry for the further 3D finite element (FE) sensitivity study.

The objectives of this paper are:

a) Evaluate the accuracy of a simple MEC model in predicting the magnetic loading and the iron losses in a 3D core;
b) Present a core loss model to calculate the losses in each element of the 3D FE core, considering the different loci of flux density variation;
c) Compare the predicted core loss with measurements in the single-phase four-pole prototype motor.

In Section II, some general remarks on the computation of the magnetic loading over the magnetization cycle are stated, and Section III deals with the loss estimation for a claw-pole structure. Section IV presents the experimental results, and Section V describes the optimization process used to find the optimal choice of material and geometry for the claw-pole motor, based on the pole number and the dimensions.

II. MAGNETIC LOADING

The specific core loss energy per revolution can be predicted on the basis of the variation and location of the magnetic loading (1), as a sum of hysteresis, classical eddy current and anomalous losses [6]-[8].

\[
w_{loss} = \int_{0}^{2\pi} \left( k_h B^2 + 2 \cdot k_e \cdot \omega \int_{0}^{2\pi} \left( \frac{dB}{d\theta} \right)^2 d\theta + 1.8 \cdot k_a \cdot \omega^{1/2} \left( \int_{0}^{2\pi} \left( \frac{dB}{d\theta} \right)^{3/2} d\theta \right)^{1/2} \right) d\theta
\] (1)

Two methods have been used to evaluate the flux density variation in different parts of a claw-pole structure, namely the MEC and FE methods, and their models are described below.

A. Magnetic Equivalent Circuit Model

The symmetric part of the outer-rotor claw-pole motor is shown in Fig. 1, where the outer magnet ring and its magnetization pattern is shown on the radial/circumferential (r-0) plane. The magnetic equivalent circuit has 18 elements, referred to as ‘e’ in (2), and they describe the main flux paths for the 3D core at the alignment position. The topology matrix (2) is used to assemble the permeance matrix G [9].

\[M_{enc} = \left[ \begin{array}{ccc} e & node_1 & node_2 \end{array} \right] G_m^e \] (2)

Fig. 1. Permeance network attached to the claw-pole structure. The outer magnet ring is at the alignment position.
The node potential method (3) is used to calculate the scalar magnetic potential, branch fluxes and magnetic loading for each element. The core elements (filled) are nonlinear and three leakage paths are considered in the core: the leakage between the claw-poles (P), between the flanks (F) and between the claw-pole and the bottom of the core (B).

\[ \mathbf{G}_u = \varphi \] (3)

In order to consider the non-zero current in the armature coil, the mmf source has to be added to the algebraic equations for the node points, as in (4).

\[ (u_3 - u_4) \cdot G_{45} + \phi_{45} = -F \cdot G_{45} \] (4)

When the rotor is located at any other position different from the alignment position (Fig. 2), there will be a flux flowing in the circumferential direction through the symmetric surfaces, described in (5),

\[ \phi_{0}(r, \theta, z_{0} + z) = -\phi_{0}(r, \theta + 2\pi/N_p, z_{0} + z) \] (5)

where \( N_p \) is the number of poles and \( z_0 \) is the coordinate of the reflection surface in the axial direction. The additional formulation that considers the node points on the periodic boundaries in the MEC model can be expressed as in (6), where the additional nodes share the same expression.

\[ \begin{bmatrix} 1 & -1 & \cdots \ & -1 \\ -1 & 1 & \cdots \ & 1 \\ \vdots & \vdots & \ddots & \vdots \\ -1 & 1 & \cdots \ & 1 \end{bmatrix} \begin{bmatrix} u_{0} \\ u_{10} \end{bmatrix} + \begin{bmatrix} u_{10} \\ u_{14} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \] (6)

The MEC model predicts the magnetic loading in four parts in the core: the base core, the flank and the two parts in the claw-pole. If the flux variation is considered only for the fundamental then the core loss per pole can be calculated. Increasing the number of elements in the MEC model to improve the accuracy also increases remarkably the permeance network formulation time.

**B. 3D Finite Element Model**

Using the FE method allows to discretize the machine in a larger number of 3D elements, which allows a more precise calculation of the magnetic loading in the machine over the excitation cycle. The model was implemented using a commercial package, Opera-3D. The equation describing the magnetization pattern in the permanent magnet is redefined in order to simulate the rotation while keeping the same 3D mesh and element numeration. This is convenient to carry out the harmonic analysis directly at each element. Alternatively, if the 3D FE mesh is rotated then the elements are renumbered, and they can only be indentified by using an external table where the original coordinates of the elements have been stored. The flow-chart of the loss calculation procedure is shown in Fig. 3.

![Fig. 3. Post-processing approach applied to the series of 3D FE magnetostatic solutions over the magnetization cycle. Opera-3D command language is used to describe the processes in the flowchart. The content of the gray block is processed outside the Opera-3D post-processor.](image)

The coordinates are calculated at the centre of the 3D FE elements (ELEM) in the database for the initial position (DBAS), before the mesh is rotated. It should be noted that when calculating the fields in the rotor, the mesh will have to be rotated back so that the rotor coordinates correspond to those at the initial position. Finally, the result table containing the calculated flux density values (RBX, RBY, RBZ), is loaded into the initial 3D FE database, and the specific core loss is calculated according to the new values at each element. The total loss is computed performing a volume integral over the FE model. The magnitudes of the fundamental flux components are shown in Fig. 4. At the alignment position the radial and the axial components are dominating, while the rotation gives rise to the circumferential component.

**III. IRON LOSS CALCULATION**

The iron losses in a rotating electrical machine consist of an alternating and a rotating component [1], [3], [7], and can be expressed as in (7). For pure alternation and rotation the trajectory of the flux density loci describes a line and a circle respectively. But in general, alternating and rotating effects interact yielding an elliptical trajectory, and \( B_{\text{major}} \) and \( B_{\text{minor}} \) represent the major and minor axis of the ellipse. Their ratio determines the contribution of the alternating and rotating components to the total core losses.

\[ \omega_{\text{core}} = \frac{B_{\text{minor}}}{B_{\text{major}}} \cdot \omega_{\text{circle}} + \left( 1 - \frac{B_{\text{minor}}}{B_{\text{major}}} \right)^2 \cdot \omega_{\text{line}} \] (7)

**Fig. 2. Permeance network at the unalignment position. The additional elements connect the corresponding symmetric surfaces.**

**Fig. 4.** The total loss is computed performing a volume integral over the FE model. The magnitudes of the fundamental flux components are shown. At the alignment position the radial and the axial components are dominating, while the rotation gives rise to the circumferential component.
When the ratio is 0 or 1 the losses are purely alternating or rotational respectively. This ratio was calculated in the claw-pole motor for the case of the circular polar magnetized ring, and a surface mounted magnet rotor, and it is shown in Fig. 5. The ratio has been calculated as in (8), where \( B_{\text{mod}} \) is the modulus of the varying flux density vector as a function of the rotor angular position.

\[
\frac{B_{\text{minor}}}{B_{\text{major}}} = \min(B_{\text{mod}}) / \max(B_{\text{mod}})
\]

(8)

It can be observed how the flux alternation occurs mainly in the base core, while the flux density loci are close to a circle in the claw-poles and an ellipse in the flanks. The advantage of the surface mounted magnet rotor is that the magnet material is reduced by 57%. However, the total harmonic distortion in the core of this machine has a maximum of 15%, compared to only 2.5% in the machine with the polar magnetized magnet ring, both delivering the same torque.

### Table I

<table>
<thead>
<tr>
<th>Loss coefficient, B locus</th>
<th>Loss coefficients</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Hysteresis, alternating B</strong></td>
<td>( k_0 = 1213 \text{ [W/m}^3\text{]}, n = 1.88 )</td>
</tr>
<tr>
<td><strong>Hysteresis, rotating B</strong></td>
<td>( a_1 = 64 \times 10^3, a_2 = 1.054, a_3 = 1.445, B_S = 2.134 \text{T} )</td>
</tr>
<tr>
<td><strong>Dynamic, alternating B</strong></td>
<td>( k_d = 57, n_d = 1.85, n_f = 1.4 )</td>
</tr>
<tr>
<td><strong>Dynamic, rotating B</strong></td>
<td>Not specified</td>
</tr>
</tbody>
</table>

The specific hysteresis core loss energy (9) depends only on the flux density locus, the peak value of \( B_{\text{mod}} \) and saturation \( B_S \). The loss coefficients used for the prediction are summarized in Table I. These coefficients have been adapted to the SMC material used in the prototype motor (Somaloy500+0.6% LB1), from the values provided in [1], [3], [7]. The specific hysteresis loss energy over the stator core is shown in Fig. 6. The picture on the left hand side shows only the alternating losses, concentrated mainly in the base core, while the picture on the right also includes the rotational losses, which are clearly concentrated in the claw-poles.

\[
\begin{align*}
W_{\text{hyst}} &= k_h \hat{B}^n \\
W_{\text{circle}} &= a_1 \left[ \frac{1}{(a_2 + 1/s)^2 + a_3^2} \right. \\
& \quad - \left. \frac{1/(2-s)}{(a_2 + 1/(2-s))^2 + a_3^2} \right] \\
S &= 1 - \frac{1}{1 - \frac{\max(B_{\text{mod}})}{B_S} \sqrt{1 - \frac{1}{a_2^2 + a_3^2}}} 
\end{align*}
\]

(9)
B. Dynamic Core Loss

The dynamic loss energy is the energy dissipated per magnetization cycle that depends on the frequency of the cycle. The dynamic core loss is the sum of the classical eddy current loss and the anomalous loss in the magnetic material. The specific dynamic power loss energy is computed according to (7).

\[ w_{\text{dyn}}^{\text{loss}} = 1.94 \cdot k_{\text{dyn}}(2\pi f)^{1.4} \sum_{i=1}^{N} \left( \Delta B_{ik}^2 + \Delta B_{ik}^2 \right)^{1.4/2} \quad (7) \]

where \( N \) indicates the number of solutions over the magnetization cycle. Equation (7) is applicable for the classical eddy current or anomalous loss calculation individually, using the appropriate coefficient and exponential terms. However, it is advantageous to use a single formulation combining all the dynamic losses, since then the loss coefficients can be calculated more easily from the measurements [6]. Even though the dynamic loss differs for the field alternation and rotation [1], [3], [7], the dynamic core loss coefficient is not calculated for the rotating fields.

IV. Core Loss in a Single-Phase Claw-Pole Motor

The prototype single-phase outer-rotor claw-pole motor has four poles with progressive radius. The difference between the gap lengths at the leading and the lagging edge of the claw-poles causes an asymmetric airgap flux distribution, and a starting torque. The magnetization pattern of the ferrite permanent magnet ring is circular polar and no back iron is used [10].

A. Core Loss Estimation

The prediction of the iron losses is based on the coefficients provided in Table I, and the results are summarized in Table II. The loss energy is given per excitation cycle. The losses of electromagnetic origin are not predicted in the permanent magnet or any other material in the vicinity of the rotating magnet ring.

B. Core Loss Measurement

A calibrated dc motor is used to estimate the energy \( W_{\text{hyst}} \) required to magnetize the core over the magnetization period, according to (8). The static friction \( T_{\text{frict}} \) is measured from the test bench without the claw-pole core attached. The time independent conservation equation (8) for the complete system includes magnetic friction (hysteresis) \( T_{\text{hyst}} \).

\[ W = \psi_{\text{dcm}} \int_0^{2\pi} i_0(\theta) d\theta = \frac{1}{2} \int_0^{2\pi} T_{\text{frict}}(\theta) d\theta + \frac{1}{2} \int_0^{2\pi} T_{\text{hyst}}(\theta) d\theta \]

\[ W = 2\pi \psi_{\text{dcm}} \int_0^{2\pi} i_0(\theta) d\theta + W_{\text{hyst}} \]

The results for the torque from the dc motor, friction and cogging are shown in Fig. 8. These components have been measured along two electric periods, which correspond to the values from 0 to 1 (first period) and from 1 to 2 (second period) in the horizontal axis in Fig. 8, where one period corresponds to 14.20 seconds.

In order to estimate the dynamic losses the rotation speed is increased gradually. The dynamic iron loss in the core \( P_{\text{dyn,core}} \) is calculated from (9). The dynamic mechanical loss \( P_{\text{dyn,mech}} \) is measured without the claw-pole core attached. Hence, neither friction between the claw-pole stator and the shaft, nor windage are included in this loss, and they will therefore contribute to overestimate \( P_{\text{dyn,core}} \). The core loss measurements are presented in Fig. 9, and they show about 35% difference between the predicted and the measured static loss and as high as 60% at 100Hz for the dynamic loss.

\[ P = 2\pi f \psi_{\text{dcm}} i_0(\theta) = 2\pi f T_{\text{frict}}(\theta) + f A_{\text{hyst}} + P_{\text{dyn,core}} + P_{\text{dyn,mech}} \quad (9) \]

<table>
<thead>
<tr>
<th>Quantity</th>
<th>Unit</th>
<th>MEC model</th>
<th>FEM model</th>
</tr>
</thead>
<tbody>
<tr>
<td>Type of outer rotor magnetization</td>
<td></td>
<td>polar</td>
<td>polar</td>
</tr>
<tr>
<td>Hysteresis loss energy at alternating</td>
<td>mJ</td>
<td>4.00</td>
<td>3.16</td>
</tr>
<tr>
<td>Hysteresis loss energy at rotating</td>
<td>mJ</td>
<td>5.82</td>
<td>5.20</td>
</tr>
<tr>
<td>Dynamic loss at alternating</td>
<td>mJ</td>
<td>0.155 ( f^{1.4} )</td>
<td>0.148 ( f^{1.4} )</td>
</tr>
</tbody>
</table>

Fig. 8. Torque no load measurement, \( T_{\text{frict}} \) without claw-pole stator, and \( T_{\text{dcm}} \) with claw-pole stator. The measurement of the cogging torque is added for comparison.

Fig. 9. Core loss and mechanic loss energy as a function of the electric rotation frequency. The additional friction between the claw-pole stator and the shaft is not considered, neither the windage loss.
V. DESIGN OPTIMIZATION

The less favorable feature of SMC material is the higher loss [4], [6], which affects the suitable domain for electromagnetic application of this material. Based on the MEC model, an optimization process has been implemented in order to find the optimal combination of size and pole number for the claw-pole motor, while keeping a constant stator volume. The results are presented in Fig. 10. The length to width ratio accounts for the different cross-sections of the claw-pole structure and the winding. The inner radius of the claw-pole ring is adjusted so that it gives a constant volume for the stator. The radial length of the rotor magnet ring and the electric loading have been kept constant. It can be observed that the highest torque values correspond to a length to width ratio close to 1. For higher pole numbers, the air gap radius – inner radius respectively, should be increased to get more torque. The magnetic loading decreases when the number of poles increases, due to the increasing leakage, and this causes lower core loss energy per excitation cycle. Also, the SMC material presents less dynamic losses at higher frequencies [4], [6], which would make it more suitable than using solid iron in the stator.

VI. CONCLUSION

The design process presented in this paper shows that the simple MEC model is sufficient to select the size of the core and to predict the magnetic loading. The flux densities, which are slightly overestimated, compared to FEM (5% error), give a loss calculation error of about 5-20% compared to FEM. A more precise loss computation can be performed using 3D FE simulations, where the loci of B can be calculated at every single element taking into account both rotational and alternating effects. It is unnecessary to put a lot of effort into accurate estimation of magnetic loading while the material characteristics for rotating fields are known only with rough approximation. According to the (low) speed controlled and calibrated dc motor, the simple loss separation between magnetical and mechanical components shows that the static magnetic loss is underestimated as much as 35% and the dynamic loss up to 60% at 100Hz.

REFERENCES


Fig. 10. Average torque of the machine as a function of the radius, core size and number of pole-pairs, while the stator volume is constant.