

Improved Analytical Model of an Outer Rotor Surface Permanent Magnet Machine for Efficiency Calculation with Thermal Effect

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ABSTRACT

In this paper, an improved analytical model accounting for thermal effects in the electromagnetic field solution as well as efficiency map calculation of an outer rotor surface permanent magnet (SPM) machine is described. The study refers in particular to an in-wheel motor designed for automotive electric powertrain. This high torque and low speed application pushes the electric machine close to its thermal boundary, which necessitates estimates of winding and magnet temperatures to update the winding resistance and magnet remanence in the efficiency calculation. An electromagnetic model based on conformal mapping is used to compute the field solution in the air gap. The slotted air-gap geometry is mapped to a simpler slotless shape, where the field solution can be obtained by solving Laplace's equation for scalar potential. The canonical slotless domain solution is mapped back to the original domain and verified with finite element model (FEM) results. Closed form solutions of core loss and magnet loss are derived from the air-gap field solution. The copper loss is calculated by considering the proximity loss and skin effects. In order to estimate the winding and magnet temperatures, a thermal model is built using a lumped parameter thermal network with an improved discretization approach. The model has been validated experimentally using the end-winding and coolant temperatures. The energy consumption calculation with the New European Driving Cycle (NEDC) is performed and the benefit of having the thermal model is quantified in terms of percentage difference in the calculated energy consumptions.

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INTRODUCTION

Electric vehicle powertrain components are optimized with a focus on minimizing the energy consumption in a driving cycle, which represents the real world vehicle operation. Analytical models, which are simple and accurate in estimating the powertrain efficiency, can reduce the computational time of the early stage design optimization process. In this paper, an analytical modelling approach for an outer rotor surface mounted permanent magnet (SPM) machine designed for an in-wheel motor electric powertrain is explained in detail.

In literature, the approaches for computing the airgap field solution of the SPM machine, which can be the basis for calculating the efficiency, are given as follows. Carter introduced the use of an equivalent air-gap that is greater than the physical air-gap to model the slotting effect [1,2] in the field solution. This method captures the overall reduction in air-gap flux density, but the local distortion of the flux density which produces cogging torque is not modelled. Zhu et al.[3] proposed a relative permeance based method to model the local radial field distortion. As this method does not model the influence of slotting in the tangential field component, Zarko et al.[4] suggested

a complex relative permeance method. The radial and tangential components of the field solution and the permeance function are represented as real and imaginary parts of a complex number, which enabled the derivation of closed form solutions for the global parameters. The harmonic model [5] and mode matching technique [6] give good accuracy with reasonable computation time. The comparison study performed in [7] indicates that the complex permeance and the mode matching techniques provide good compromise between accuracy and computational time. The main missing topic to be addressed is the study of temperature influence in the efficiency map calculation.

In order to resolve the above mentioned problem, a thermal model based on lumped parameter thermal networks is built using the Distributed Loss & Capacitance (DLC) element as proposed in [8]. The network is solved by introducing the losses obtained from the electromagnetic model and considering the effect of rotational speed on some of the boundary conditions. The temperature affects the properties of magnet and winding in the machine. Radially magnetized neodymium magnet is used in the SPM machine and its remanence reduces with increasing temperature until a maximum allowable point, above which demagnetization of the magnet occurs. On the other hand, the temperature rise increases the DC loss contribution in the total winding loss but reduces the AC loss [9]. The coupled electromagnetic-thermal model discussed in this paper accounts for these effects allowing efficient, due to the reduced computational effort, and improved estimation of the machine losses and performance.

1. MATHEMATICAL MODELLING OF ELECTRIC MACHINE

The air gap field solution of the machine, which is the basis for calculating the global parameters such as torque, losses, and efficiency, is obtained through a conformal mapping procedure. The saturation effect is not included in the calculation based on the assumption that the iron is infinitely permeable. The outer rotor surface permanent magnet machine with radial magnetization developed for automotive application is shown in Figure 1 and its cross-sectional view is given in Figure 2.



Figure 1. In-wheel motor assembled in the vehicle.

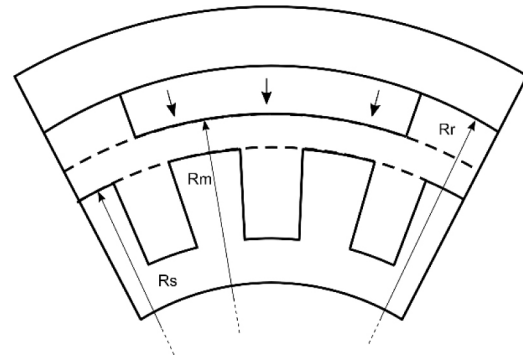


Figure 2. Single pole pitch cross section of outer rotor SPM motor.

The motor parameters are given in Table 1. However, due to confidentiality reasons, the dimensions of the motor are not given in the paper

Table 1. Parameters of external rotor surface PM motor

Continuous maximum power	52 kW
Continuous maximum torque	500 Nm
Rated voltage	320 V
Rated current	150 A
Number of poles	40
Number of slot	120
Number of conductor	8

1.1 Electromagnetic Torque Calculation

The field produced by the permanent magnets in the slotted air gap (B_s) is computed from the slotless air-gap field solution (B_k) and complex relative air-gap permeance (λ) as in (1). The term λ accounts for slotting effect in the field solution [10, 11].

$$B_s = B_k \lambda^* \quad (1)$$

$$B_{sr} + jB_{s\theta} = (B_{kr}\lambda_a + B_{k\theta}\lambda_b) + j(B_{k\theta}\lambda_a - B_{kr}\lambda_b) \quad (2)$$

where r and θ denotes the radial and tangential components of the field solutions. Similarly, λ_a and λ_b in (2) capture the slotting effect in the radial and tangential directions.

Conformal transformation, a special kind of complex function that preserves the angle between intersecting curves at all points in the mapped complex variable planes, is employed as the perpendicularity between equipotential lines and flux paths should be preserved for magnetostatic calculations. The slotted air gap shape (S plane) is transformed into a slotless air gap (K plane) using four different conformal mapping functions as shown in Figure 3.

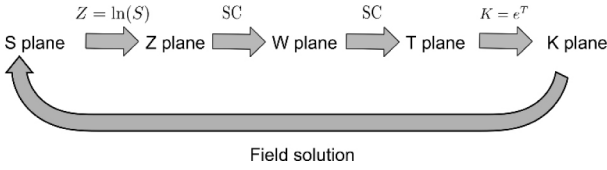


Figure 3. Analytical conformal mapping sequence.

The closed form field solution of slotless air gap field was derived by Zhu et al. in [10].

$$B_{str} = \sum_{n=1,3,\dots}^{\infty} \frac{(-4npB_r \sin \frac{n\pi\alpha_p}{2})}{n\pi\mu_r((np)^2 - 1)} \left[\left(\frac{r}{R_m}\right)^{np-1} + \left(\frac{R_s}{R_m}\right)^{np-1} \left(\frac{R_s}{r}\right)^{np+1} \right]$$

$$\left\{ \frac{(np-1)\left(\frac{R_m}{R_r}\right)^{2np} + 2\left(\frac{R_m}{R_r}\right)^{np-1} - (np+1)}{\frac{\mu_r+1}{\mu_r}\left[1 - \left(\frac{R_s}{R_r}\right)^{2np}\right] - \frac{\mu_r-1}{\mu_r}\left[\left(\frac{R_s}{R_m}\right)^{2np} - \left(\frac{R_m}{R_r}\right)^{2np}\right]} \right\} \cos(np\theta)$$

$$B_{sl\theta} = \sum_{n=1,3,\dots}^{\infty} \frac{(-4npB_r \sin \frac{n\pi\alpha_p}{2})}{n\pi\mu_r((np)^2 - 1)} \left[-\left(\frac{r}{R_m}\right)^{np-1} + \left(\frac{R_s}{R_m}\right)^{np-1} \left(\frac{R_s}{r}\right)^{np+1} \right]$$

$$\left\{ \frac{(np-1)\left(\frac{R_m}{R_r}\right)^{2np} + 2\left(\frac{R_m}{R_r}\right)^{np-1} - (np+1)}{\frac{\mu_r+1}{\mu_r}\left[1 - \left(\frac{R_s}{R_r}\right)^{2np}\right] - \frac{\mu_r-1}{\mu_r}\left[\left(\frac{R_s}{R_m}\right)^{2np} - \left(\frac{R_m}{R_r}\right)^{2np}\right]} \right\} \sin(np\theta)$$

(3)

and where $np=1$,

$$B_{str} = \sum_{n=1,3,\dots}^{\infty} \frac{2npB_r \sin \frac{n\pi\alpha_p}{2}}{n\pi\mu_r} \left[1 + \left(\frac{R_s}{r}\right)^2 \right]$$

$$\left\{ \frac{\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_s}\right)^2 + \left(\frac{R_r}{R_s}\right)^2 \ln \left(\frac{R_m}{R_r}\right)^2}{\frac{\mu_r+1}{\mu_r}\left[1 - \left(\frac{R_r}{R_s}\right)^2\right] - \frac{\mu_r-1}{\mu_r}\left[\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_m}\right)^2\right]} \right\} \cos(\theta)$$

$$B_{sl\theta} = \sum_{n=1,3,\dots}^{\infty} \frac{2npB_r \sin \frac{n\pi\alpha_p}{2}}{n\pi\mu_r} \left[-1 + \left(\frac{R_s}{r}\right)^2 \right]$$

$$\left\{ \frac{\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_s}\right)^2 + \left(\frac{R_r}{R_s}\right)^2 \ln \left(\frac{R_m}{R_r}\right)^2}{\frac{\mu_r+1}{\mu_r}\left[1 - \left(\frac{R_r}{R_s}\right)^2\right] - \frac{\mu_r-1}{\mu_r}\left[\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_m}\right)^2\right]} \right\} \sin(\theta)$$

(4)

The magnet temperature affects the remanence (B_r) in (3), (4) as well as the resulting field solution. The complex relative air-gap permeance introduced for an inner rotor machine using conformal mapping by Zarko et al in [11] is recalculated for the outer rotor machine and the generic expression is given as,

$$\lambda^* = \left(\frac{\partial k}{\partial s}\right)^* = \left(\frac{\partial k}{\partial t} \frac{\partial t}{\partial w} \frac{\partial w}{\partial z} \frac{\partial z}{\partial s}\right)^*$$

(5)

Each partial derivative in (5) is defined by the conformal transformation shown in Figure 3 and the final expression of the complex relative permeance [11] can be written as

$$\lambda = \lambda_a + j\lambda_b = \left(\frac{k}{s} \frac{(w-1)}{(w-a)^{\frac{1}{2}}(w-b)^{\frac{1}{2}}} \right)$$

(6)

The assumptions made in this method is that the evaluation points form a circular arc in the canonical domain and the magnets retain their shape. Although these assumptions enable the derivation of the closed form solutions of global parameters, they impair the accuracy of the tangential component of field solution, in Figure 5, and the cogging torque. However, the mean torque value obtained from this method is accurate as the radial component of the field solution confirms the FEM result as shown in Figure 4.

The flux linkage (Ψ_{pm-ph}) of the phase winding is obtained by integrating the radial flux density across one coil pitch and multiplying it with the number of coils in series (N_s) [4]. The distribution factor and pitch factor of the winding scheme for this particular motor are equal to one.

$$\Psi_{pm-ph} = N_s l_a r \int_{-\frac{\gamma_c}{2}}^{+\frac{\gamma_c}{2}} B_{sr}(r, \theta, t) d\theta$$

(7)

The stator flux linkage in the d axis (Ψ_d) is the sum of permanent magnet flux linkage in d axis (ψ_{pm-d}), which is derived from Park transformation of the flux linkages in three phase windings, and the contribution from the d -axis current.

$$\psi_d = \psi_{pm-d} + L_d I_d$$

(8)

In the surface mounted permanent magnet machine, the inductance variation in the d - q axes is negligible ($L_d = L_q = L_s$), so the electromagnetic torque is computed from the flux linkage in the d axis (Ψ_d) and the stator current in q axis (I_q) [12].

$$T_{em} = \frac{3p}{2} \psi_d I_q$$

(9)

For the purpose of efficiency calculation, the I_q current required to produce the demanded output torque is calculated from (9). In the field weakening region the I_d current is supplied based on the operating speed of the motor as derived in (10) [12].

$$I_d = \frac{\psi_d^2 + L_s^2 I_{max}^2}{2L_s \Psi_d} \left[\left(\frac{\omega_s^A}{\omega} \right)^2 - 1 \right]$$

(10)

where L_s is the stator inductance and I_{max} is the maximum stator current. The maximum voltage (V_s^{max}) defines the rated speed of the motor ω_s^A as shown in Figure 6.

$$\omega_s^A = V_s^{max} / \sqrt{(L_s I_q)^2 + (\psi_{pm} + L_s I_d)^2} \quad (11)$$

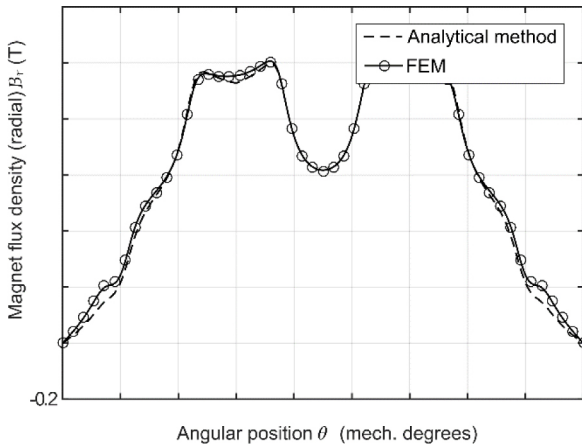


Figure 4. Comparison of radial flux density with FEM result

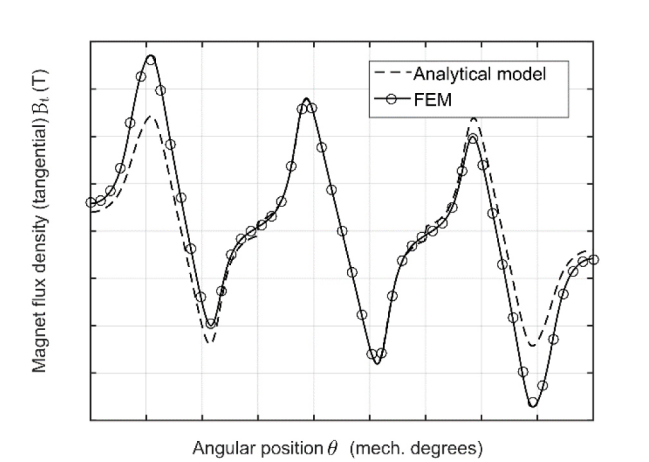


Figure 5. Comparison of tangential flux density with FEM result

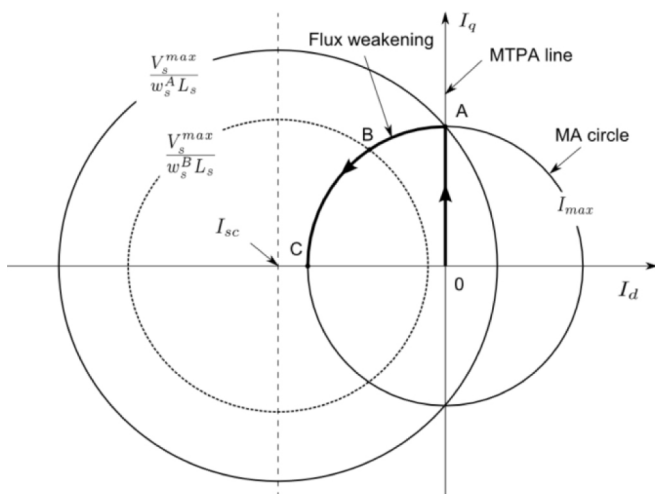


Figure 6. Current locus diagram of outer rotor surface permanent magnet machine with short-circuit current (I_{sc}) greater than maximum allowable stator current (I_{max})

1.2 Efficiency Calculation

The efficiency of the motor is defined as the ratio between output power at the shaft (P_{out}) and the input electrical power(P_{in})

$$\eta = \frac{P_{out}}{P_{in}} \quad (12)$$

The output power is calculated by subtracting the core losses (P_c), magnet losses (P_{pm}) and mechanical losses (P_{ml}) from the electromagnetic power (P_{em}).

$$P_{out} = P_{em} - P_c - P_{pm} - P_{ml} \quad (13)$$

The input power (P_{in}) is calculated by adding the armature loss (P_a) with the electromagnetic power (P_{em}).

$$P_{in} = P_{em} + P_a \quad (14)$$

1.2.1 Copper Losses

The copper loss model includes AC losses due to skin effect inside conductors and additional losses due to proximity effect. It also captures the effect of winding temperature on the losses.

$$P_a = 3 I_s^2 R_{AC} + P_{proximity} \quad (15)$$

The effective resistance (R_{AC}) that accounts for the skin effect in the conductor is modelled as a function of conductor dimensions, operating frequency and winding temperature. The conductors are layered in the slot as shown in [Figure 7](#). The detailed explanation of the model can be found in [\[13\]](#).

$$R_{AC} = K_{AC} N_s \frac{1}{\sigma_0} \frac{2(l_a + l_{ew})}{A_c} (1 + \beta_t(T - T_0)) \quad (16)$$

The AC loss factor K_{AC} can be modelled as,

$$K_{AC} = \left(\frac{360+3\gamma^4}{360+\gamma^4} + \left(\frac{\sum_{l=1}^N NI}{N_l I_0} \right)^2 \frac{2\gamma^4}{24+\gamma^4} \right)$$

where $\Sigma_{l=1}^N NI$ and N_l are defined according to Figure 7 [13], I_0 is the current in layer l , N_s is the number coils connected in series, σ_0 is the conductivity of copper at reference temperature ($T_0 = 20^\circ C$), l_a is the stack length, l_{ew} is the end winding length, A_c is the cross sectional area of the wire, and T is the winding temperature. The characteristic dimension γ and the AC loss temperature coefficient β , are defined as

$$\gamma = 0.93 \sqrt{\epsilon_l} \frac{D_c}{\delta} \quad (17)$$

$$\beta_t = \alpha_t - \alpha_t \frac{1}{k_{AC}} \left(\frac{\sum_{l=1}^n NI}{NI_0} \right)^2 \frac{384\gamma^4}{(24 + \gamma^4)^2}$$

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