

Optimal Design of an Interior Permanent Magnet Synchronous Motor for Wide Constant-Power Region Operation: Considering Thermal and Electromagnetic Aspects

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ABSTRACT

The paper proposes a design optimisation of an Interior Permanent Magnet synchronous motors with maximum output power density and suitable for wide constant-power region operation. In this paper, analytical magnetic and electrical models of the machine are developed to calculate parameters and variables of the machine needed for a design optimization such as flux, resistance and inductances. And then, the thermal aspect is modelled using a thermal lumped-parameter network which allows to estimate the machine temperatures at key points such as the windings and the magnet. These models are included in the optimization loop and so are evaluated at each iteration. The optimization method uses a differential evolution algorithm (DEA). Finally, output performances of the designed motor are verified by finite element analysis (FEA).

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INTRODUCTION

The Interior Permanent Magnet (IPM) synchronous motors, built with magnets placed inside the rotor body, are attracting great attention in several variable speed applications, such as electric vehicles, industrial and domestic appliances, where the most challenging requirements are high efficiency, high torque density, good overload capability and extended speed range. Additional features are the robustness of the rotor structure, mechanically suited to high speed operation, and the presence of magnetic saliency: the "direct" d -axis inductance is substantially different from the "quadrature" q -axis inductance, where the d -axis is aligned with the Permanent Magnet (PM) flux according to the equivalent Park model of the synchronous machine. This characteristic is particularly suited for extending the torque/speed operating region by proper "field weakening" control techniques: the most popular approach is to combine the maximum torque per ampere trajectory with the voltage-constraint-tracking field-weakening control 1-5. The objective to improve the flux-weakening capability of IPM motors finds a wide interest in the literature. This paper proposes a design optimization procedure of Interior Permanent Magnet synchronous motors suitable for wide constant-power region operation with the aim to maximize the power density taking into account the electromagnetic and thermal aspects.

Peculiarities of the proposed approach are the following:

1. The optimization is performed to meet the typical specifications for a hybrid (or electric) vehicle application taking into account the electromagnetic and thermal aspects.
 - Nominal torque at base speed
 - Nominal torque at high speed
 - Maximum power at base speed for 3 min
 - Maximum power at high speed for 3 min.
2. the design concerns all the motor (stator and rotor core, winding and magnets);
3. some of the design variables vary in a discrete way with fixed steps;
4. The design procedure does not need to start from a known initial design, i.e. a feasible initial design is not required.

The proposed design optimization of IPM motors is based on a new differential Evolution (DE) optimization algorithm proposed in [2..]. The analytical models are used.

1. THE DESIGN APPROACH

To guarantee a wide constant power operation, the motor requires an accurate design through use of salient rotor

geometry with limited flux contribution from PMs buried within the rotor structure. To achieve the desired degree of saliency, maximize the power density and guarantee good performance, special lamination profile should be found. The particular rotor geometry of the IPM synchronous motor and the presence of several variables and constraints make the design problem very complicated to solve. This paper proposes a new design approach of the IPM synchronous motor allowing to guarantee a desired wide constant power operation. This approach is based on combining the analytical models to with mathematical optimization algorithms. The analytical models are developed to take account of the electromagnetic and thermal aspects in the design process.

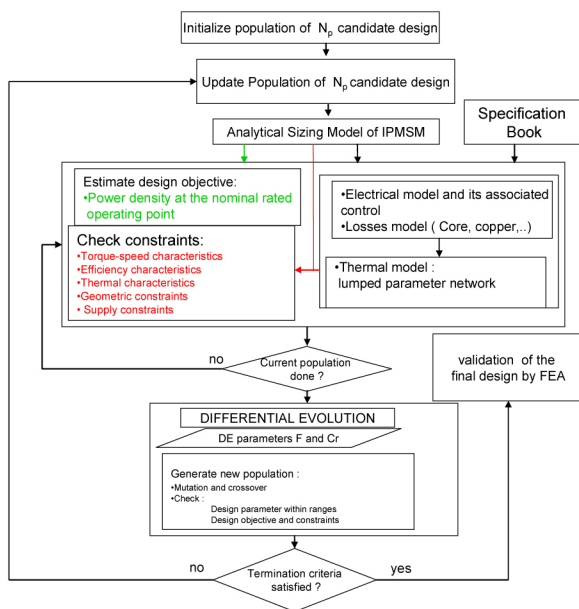


Figure 1. Design optimization process of IPM motor employing analytical models and ICDE optimizer

1.1. Differential Evolution (DE) Optimization Algorithm

In this paper a novel DE method proposed by 8 for constrained optimization problems “named ICDE” has been used to design IPM motor. ICDE is an improved version of $(\mu+\lambda)$ -constrained differential evolution ($(\mu+\lambda)$ -CDE) proposed by 9. ICDE mainly consists of an improved $(\mu+\lambda)$ -differential evolution (IDE) and a novel archiving-based adaptive tradeoff model (ArATM). Therein, IDE employs several mutation strategies and the binomial crossover of differential evolution (DE) to generate the offspring population. Moreover, In ICDE a new mutation strategy named “current-to-rand/best/1” is proposed by making use of the current generation number in IDE. Since the population may undergo three situations during the evolution (i.e., the infeasible situation, the semi-feasible situation, and the feasible situation), like $(\mu+\lambda)$ -CDE, ArATM designs one constraint-handling mechanism for each situation. However, unlike $(\mu+\lambda)$ -CDE, in the constraint-handling mechanism of the infeasible situation, the hierarchical non dominated individual selection scheme is utilized, and an individual archiving technique is proposed by to maintain the diversity of the

population. Furthermore, in the constraint-handling mechanism of the semi-infeasible situation, the feasibility proportion of the combined population consisting of the parent population and the offspring population is used to convert the objective function of each individual. It is noteworthy that ICDE adopts a fixed tolerance value for the equality constraints. In addition, in ICDE two criteria are used to compute the degree of constraint violation of each individual in the population, according to the difference among the violations of different constraints. By combining IDE with ArATM, ICDE has the capability to maintain a good balance between the diversity and the convergence of the population during the evolution. The performance of ICDE has been tested on 24 well-known benchmark test functions collected for the special session on constrained real-parameter optimization of the 2006 IEEE Congress on Evolutionary Computation (IEEE CEC2006). The experimental results demonstrate that ICDE not only overcomes the main drawbacks of $(\mu+\lambda)$ -CDE but also obtains very competitive performance compared with other state-of-the-art methods.

1.2. Analytical Sizing Model of the IPM Motor

The analytical sizing model of the IMP presented here is constituted of relations which translate the physical behavior of the structure (electromechanical conversion, flux conservation) and the basic machine geometrical constraints. Reference 6 presents extensive aspects of PM motor technology and design. Some of these concepts are used in this paper. Using the design variables as input to this model, the geometrical parameters of the IPM motor can be obtained. The geometrical characteristics of the IPM motor are shown in Figure 2.

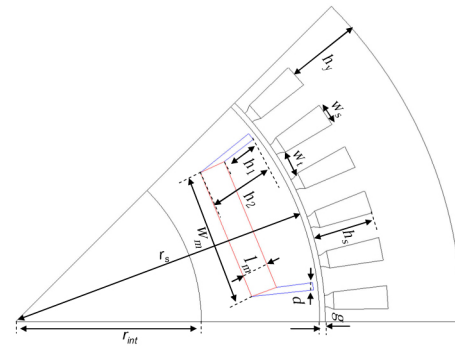


Figure 2. The geometrical characteristics of the IPM motor

Stator Design

The stator yoke thickness is obtained as follows:

$$h_y = \frac{B_{mg} \cdot \tau_p}{2 \cdot k_f \cdot B_s} \quad (1)$$

where $B_{mg} = B_g \cdot \alpha_i$ represents the average air gap induction, τ_p the pole pitch, k_f the axial iron percentage and B_s the stator yoke induction. Thus a tooth's width w_t is computed in function of the tooth induction B_d using the flux conservation law.

$$w_t = \frac{B_g \cdot \tau_s}{k_f \cdot B_d} \quad (2)$$

where τ_s is the slot pitch.

The slot widths are then obtained from the bore radius and the number of slots per pole per phase N_{spp} :

$$w_s = \tau_s - w_t \quad (3)$$

The slot surface is determined using the following equation:

$$A_s = \frac{\hat{A}_t \tau_s}{J_c \cdot K_r} \quad (4)$$

where K_r is the slot fill factor and J_c the surface current density. and the slot depth d_s is given by:

$$d_s = \frac{A_s}{w_s} \quad (5)$$

Magnet sizing:

A detailed magnetic equivalent circuit based on the geometrical structure of the motor [Figure 2](#) can be used to calculate the average air gap flux density as:

$$B_g = \frac{C_\Phi}{1 + \beta(1 + 2\eta + 4\lambda)} B_r \quad (6)$$

The magnetic reluctances of stator and rotor cores are ignored. The values of parameters in (6) are given by:

$$\beta = \frac{\mu_{rec} K_c g}{w_m} C_\Phi \quad (7)$$

$$\eta = \frac{w_m (h_1 + h_2)}{4 d \mu_{rec} l_m} \quad (8)$$

$$\lambda = \frac{1 + 1/\beta + 2\eta}{2(A_m / A_{mm})(B_r / B_s) - 4} \quad (9)$$

where B_r is the remanence of the magnet, $C_\Phi = A_m / A_g$ the flux concentration factor and A_g and A_m are the cross-sectional areas per pole of the air gap and magnet, respectively. where $A_g = \alpha r_s l m / p$, $A_m = w_m l r$, α is the pole-arc to pole-pitch ratio and

k_c is the Carter coefficient, g is the air gap length, μ_{rec} the recoil permeability and $A_{mm} = t l_r$ represents the cross sectional area of the iron bridge above the nonmagnetic barriers with t and l_r being the bridge width and motor stack length, respectively. Also l_m and w_m denote the magnet length and width; and h_1 and h_2 represent the inner and the outer flux barrier heights, respectively. B_s is a limit of the leakage flux density in the bridge due to saturation.

Combining (6), (7), (8), (9), the magnet length can be expressed as:

$$l_m = \frac{\left(\frac{1}{2} (h_1 + h_2) \cdot k_c \cdot g \cdot p + \pi \cdot \alpha \cdot r_s \cdot d \right) \cdot B_{g1} + 2 \cdot d \cdot p \cdot t \cdot B_s}{d \cdot p \cdot (l_m \cdot B_r - B_{g1} \cdot \mu_{rec} \cdot k_c \cdot g)} \quad (10)$$

Calculation of the Electromagnetic Parameters

Electromagnetic parameters of the motor are computed from the previous geometric variables.

The magnetic flux Φ_s and the stator resistance R_s are defined as follows:

$$\Phi_s = 2 k_{lw} N_{spp} r l_r B_g N_{cs} \quad (11)$$

$$R_s = 2 \rho_{cuivre} \left[l_r + \frac{\pi^2 (r_s + 0.5 d_s)}{2 p} \right] \frac{p N_s}{\pi r_s d_s K_R} N_{cs}^2 \quad (12)$$

The d - and q -axis inductances can be expressed as 10

$$L_d = \frac{6 \mu_0 r_s l_r}{\pi g_d} k_{w1}^2 N_{spp}^2 \cdot N_{cs}^2 + L_f \quad (13)$$

$$L_q = \frac{6 \mu_0 r_s l_r}{\pi g_q} k_{w1}^2 N_{spp}^2 \cdot N_{cs}^2 + L_f \quad (14)$$

$$g_d = \frac{k_c g}{k_{lad} - (k_l k_{ad} / (1 + \beta(1 + 2\eta + 4\lambda)))}$$

$$g_q = \frac{k_c g}{k_{lad}} \quad (15)$$

with

$$k_{lad} = \alpha + \frac{\sin \alpha \pi}{\pi}, \quad k_{laq} = \alpha + 0.1 + \frac{\sin 0.1 \pi - \sin \alpha \pi}{\pi},$$

$$k_{ad} = \frac{\sin(\alpha \pi / 2)}{\alpha \pi / 2}$$

