

Influence of the Coil Pitch and the Slot/Pole Number Combination upon the Performance of Permanent Magnet Motors

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ABSTRACT

This paper presents a mathematical investigation of the influence of the slot/pole number combination on the iron loss of permanent magnet (PM) motors. A simplified electromagnetic model of PM motors was used to develop a mathematical method of evaluating iron loss for any combination of slots and pole pairs. An investigation of the magnetomotive force distribution of stator teeth and its expression as a complex Fourier series expansion revealed that the coordinate system can be easily transformed, thereby enabling rotor iron loss to be calculated. A core factor was defined on the basis of the calculated iron losses and a map of slot/pole number combinations was created. Several promising combinations were selected from the map and their respective advantages and disadvantages were identified. A new promising combination was found featuring windings with a coil pitch of two slots. It is noteworthy that this slot/pole number combination has characteristics in-between concentrated windings and distributed windings.

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1. INTRODUCTION

Permanent magnet (PM) motors have played a key role in reducing the size and improving the efficiency of power sources for electrified vehicles in recent years. The motors used on electrified vehicles serve various purposes. First, they are installed between an internal combustion engine and the transmission on hybrid vehicles to function as a drive motor. Because of the limited space available in the engine compartment, it is necessary to downsize the motor. Second, they are also used on electric vehicles (EVs). In-wheel motors have been proposed and researched for EVs in recent years with the aim of further enhancing the value of these vehicles [1],[2],[3]. Permanent magnet motors for in-wheel drive systems must also be reduced in size to avoid interference with suspension members.

Because of these application requirements, PM motors must be further reduced in size. In order to downsize a PM motor, it is necessary to adopt a high reduction ratio, but that results in higher motor operating speeds. This gives rise to a problem of greater iron loss owing to the resultant faster flux density fluctuations in the core. Accordingly, iron loss represents a critical issue for motors used on electrified vehicles today. Because it is difficult to remove heat from the

rotor, rotor iron loss is an especially important design issue for motors that have to operate at high speed for long periods of time on electrified vehicles.

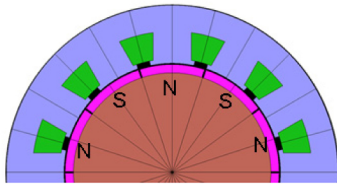
Research on slot/pole number combinations has been conducted at various research institutes over many years for application to induction motors and other electrical machines [4],[5],[6],[7]. A technique exists for reducing the copper loss and vibration of motors by applying a suitable slot/pole number combination. The authors have applied an optimal slot/pole number combination strategy to develop a coaxial motor with two output shafts and driven by compound current [8] and an axial gap motor with high torque density [9]. The use of multi-phase power supplies has achieved more degrees of freedom for motor motion, higher torque densities (i.e., lower copper losses) and lower levels of vibration.

However, there are no reports in the literature of studies that have comprehensively examined the influence of the slot/pole number combination on iron loss, including a consideration of rotor iron loss. For example, Salminen et al. mathematically derived the stator iron loss from the distribution of the magnetomotive force (MMF) harmonics of fractional-slot PM motors, which have the fractional value of q (number of slots per pole per phase). However, the assessment of rotor iron loss was limited to eddy current

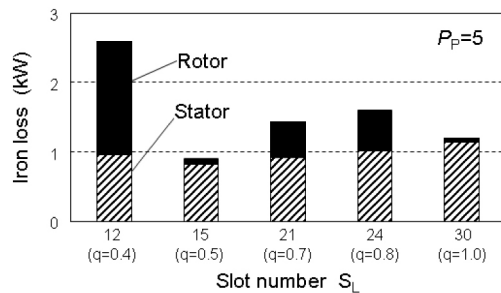
losses in the magnets caused by pulse width modulation (PWM) harmonics [5].

There is also the problem that the iron loss of fractional-slot motors becomes larger. Figure 1 compares the iron loss calculated for ten-pole (pole pair number $P_p = 5$) motors for various slot numbers S_L under identical operating conditions using the finite element method (FEM) and the JMAG ver. 9.0 electromagnetic field analysis software. All of these motors have the same size (200mm dia., 135mm stacking length) and the same base speed (5,000rpm). The results indicate that the asynchronous MMF harmonic component increases especially for 12 slots ($q = 0.4$), causing larger iron loss.

In order to calculate rotor iron loss, the spatial distribution of MMF and its transformation to the rotor coordinate system is necessary, which complicates any mathematical investigation. Therefore, a method is proposed here for facilitating a transformation to the rotor coordinate system by assuming a simple function to express the MMF spatial distribution of one stator tooth and performing a complex Fourier series expansion. This method was used in this work to seek a slot/pole number combination for reducing iron loss, including rotor iron loss. Among the various combinations examined, several promising slot/pole number combinations were selected and their advantages and disadvantages were investigated.



(a). FEM model ($S_L=12$, $P_p=5$, Stator dia. =200mm, Rotor dia. =130mm, Stacking length =135mm)



(b). Calculation results (at 10,000rpm, 150kW)

Figure 1. Influence of slot number on iron loss

2. FEATURES OF PROPOSED METHOD

The output torque of a motor is proportional to the winding factor k_W and the current [9]. The copper loss L_{COPPER} varies as the square of the current. Thus, the effect of the slot/pole number combination on L_{COPPER} can be

expressed as shown in Eq. (1) using k_W as an index. k_W is uniquely determined by the four parameters of slot/pole number combination, i.e., S_L , P_p , coil pitch C_p and number of phases M .

$$L_{COPPER} \propto 1/k_W^2 \quad (1)$$

Iron loss L_{IRON} can also be uniquely determined based on the slot/pole number combination. A core factor k_C was defined as shown in Eq. (2) as an index independent of the motor geometry, operating conditions and other aspects. A calculation of k_C will be undertaken later in Section 3.

$$L_{IRON} \propto 1/k_C^2 \quad (2)$$

The proposed method facilitates a mathematical investigation using a magnetic circuit model that is simplified to the smallest possible mathematizable range while still taking into account divergence from an actual motor. The investigation process and the assumptions made regarding model simplification are summarized in Table 1. The most significant feature of this method is that it facilitates a transformation to the rotor coordinate system as noted in (D) in the table. This is accomplished by assuming a simple function to express the MMF spatial distribution of one tooth and performing a complex Fourier series expansion as noted under (A) in the table.

3. PROPOSED METHOD

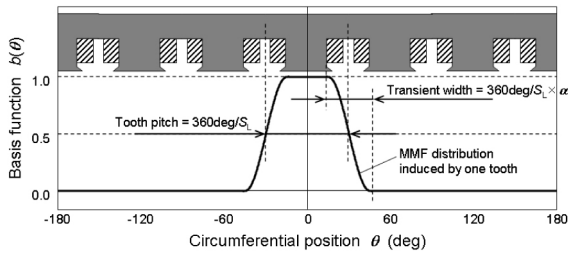
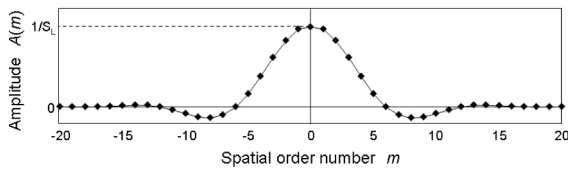
3.1. INVESTIGATION OF MMF DISTRIBUTION OF STATOR TOOTH

First, we will investigate the spatial distribution of the MMF generated by one tooth. If the MMF distribution induced by one tooth in the circumferential direction is treated as a simple pulse pattern, the sharp MMF fluctuation at the end of the pulse would increase the flux density fluctuation of the rotor, resulting in a substantial increase in rotor iron loss that would deviate from an actual motor. Therefore, the spatial distribution of the MMF generated by one tooth is expressed as a basis function $b(\theta)$ that combines a square wave and a sinusoidal wave as shown in Fig. 2(a). This can be expressed by the following equation.

$$b(\theta) = \begin{cases} 1 & \left| \frac{S_L}{\pi} \theta \right| \leq 1 - \alpha \\ \frac{1}{2} \cos \left\{ \frac{S_L}{2\alpha} \left(\theta + \pi(\alpha - 1) \right) \right\} + \frac{1}{2} & 1 - \alpha < \left| \frac{S_L}{\pi} \theta \right| \leq 1 + \alpha \\ 0 & \text{otherwise else} \end{cases} \quad (3)$$

Table 1. Process and assumptions for proposed method

Process	Description and assumptions
(A) Investigation of MMF distribution of one tooth (subsection 3.1)	Description: MMF distribution of one tooth is expressed by a complex Fourier series expansion. Assumption: The spatial distribution of MMF generated by one tooth has the combined shape of a square wave and a sinusoidal wave.
(B) Investigation of air-gap flux density (subsection 3.2)	Description: Express the air-gap flux density using the complex Fourier series expansion in (1). Assumptions: The electromagnetic circuit is represented by a 2D model. Armature MMF distribution is sinusoidal. No salient poles for the rotor and stator (magnetic reluctance distribution in the circumferential direction). Magnetic reluctance is present only in the air gap. Magnetic reluctance of the iron core is zero.
(C) Nondimensionalization (subsection 3.3)	Description: Nondimensionalize the results of (2) so as to remove the effects of motor size and operating conditions.
(D) Investigation of rotor coordinate system (subsection 3.4)	Description: Transformation to the rotor coordinate system to facilitate investigation of rotor iron loss.
(E) Investigation of iron loss (subsection 3.5)	Description: Calculate nondimensionalized iron loss from the results obtained in (3) and (4). Assumption: Only eddy current loss is considered as iron loss. Volume generating iron loss is inversely proportional to the spatial order number of the magnetic flux density.

**(a). Basis function for tooth MMF****(b). Spectrum of basis function****Figure 2. Example of basis function and its spatial spectrum ($S_L=6$, $\alpha=0.5$)**

where α is a transient width ratio that has a value in the range of $0 < \alpha < 1$. In examining the influence of the slot/pole number combination on iron loss, a major question is what value should be given to α . Calculations were performed with several different values of α , and the resultant tendency did not show any appreciable difference for $\alpha > 0.3$. Based on the calculation results, the core factor was calculated in this study using a representative value of $\alpha=1$.

The basis function expressed in Eq. (3) was subjected to a complex Fourier series expansion to obtain the amplitude of each order component $A(m)$ in order to simplify the transformation to the rotor coordinate system that will be explained in subsection 3.4. The letter m denotes a spatial order number ($-\infty < m < +\infty$). An example of the basis function spectrum is shown in Fig. 2(b).

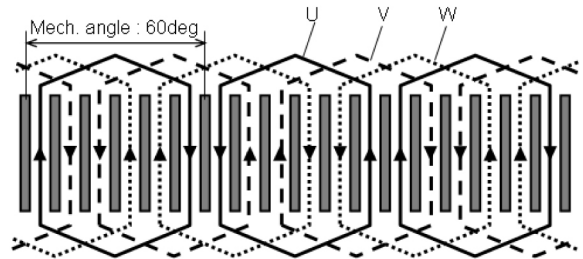
$$A(m) = \frac{1}{m\pi(1 - 4m^2\alpha^2/S_L^2)} \sin \frac{m\pi}{S_L} \cos \frac{m\pi\alpha}{S_L} \quad (4)$$

At $m = 0$ and $S_L/2/\alpha$, it is impossible to define each order component $A(m)$ because the denominator is zero. However, a limiting value exists, and that value is used here. Using Eq. (4), the basis function can be rewritten as shown below:

$$b(\theta) = \sum_{m=-\infty}^{+\infty} \frac{\sin \frac{m\pi}{S_L} \cos \frac{m\pi\alpha}{S_L}}{m\pi(1 - 4m^2\alpha^2/S_L^2)} \exp(-jm\theta) \quad (5)$$

3.2. INVESTIGATION OF AIR-GAP FLUX DENSITY

Next, we will investigate the air-gap flux density using the basis function shown above. In an investigation of a single-layered winding using a 2-D model, the same electromagnetic circuit can be expressed in a double-layered winding. In order to avoid increasing the number of patterns examined in the mathematical investigation, a single-layered winding is treated here in terms of an equivalent double-layered winding. For example, a 6-pole, 36-slot, single-layered 3-phase winding having a coil pitch of five slots (Fig. 3(a)) can be expressed in a 2-D model as being equivalent to a 6-pole, 36-slot, double-layered 3-phase winding having a coil pitch of six slots (Fig. 3(b)).

**(a). Single-layered winding****Figure 3. Equivalent expression of single-layered winding**