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Design and Analysis of a New Outer-Rotor Permanent-Magnet Flux-Switching Machine for Electric Vehicle

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Abstract

Purpose − The purpose of this paper is to propose a new outer-rotor permanent-magnet flux-switching machine for in-wheel electric vehicle propulsion. The paper documents both the design procedure and performance investigation of this novel machine.

Design/methodology/approach − The topology and preliminary sizing equations of the outer-rotor permanent-magnet flux-switching machine are introduced. Both the number and width of rotor poles are then optimized using comprehensive two-dimensional FEA. The machine losses are particularly investigated by transient FEA for the optimal design.

Findings − A outer-rotor permanent-magnet flux-switching machine, with 12 stator poles and 22 rotor poles, is most suitable for the proposed application. The analytical sizing equations are quite efficient with a suitable level of accuracy for preliminary design. The optimal rotor pole width from the FEA results is nearly 1.3 times of the original one. The efficiency of the proposed machine under rated load is relatively low, nearly 85%, as a result of significant eddy current losses in the permanent magnets, which can be effectively suppressed by implementing segmentation. The predicted outstanding performance implies that by adopting magnet segmentation the proposed machine is a leading contender for direct electric vehicle drives.

Research limitations/implications − The end effects, which could be considerable in the machine with relatively short axial length, are neglected during the study. In addition, due to the high current density and deep slot, proximity losses in the winding which is not issued in this research could be significant. All the limitations mentioned above could bring corresponding errors to the results. Although the research is concentrated on the application of electric vehicle drive, the techniques can be potentially employed for other applications.

Practical implications − The practical implementation of such a machine is confronted with several mechanical hurdles, especially the thermal issues which can be addressed by implementing innovative cooling system.

Originality/value − The outer-rotor permanent-magnet flux-switching machines so far have not been addressed yet. This research provides designers with the technical background and another alternative for electric vehicle propulsion.

Keywords Electric vehicle propulsion, Finite element analysis, In-wheel direct drive, Outer-rotor permanent-magnet flux-switching machines

Paper type Research paper

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1. Introduction

Due to drastic issues on the protection of the environment and the shortage of fossil fuels, political and public pressures helped a lot to develop practical and efficient electric vehicles (EVs) during this decade. It is fully convinced that the zero-emission electric vehicles will be rapidly exposed on and dominate the future automotive market. The propulsion system has been one of the most essential parts in EV, direct drive EVs which are propelled by in-wheel or wheel electric machines without differential gears have drawn significant attentions in both industrial and academic researches. Direct drive propulsion systems can enjoy their implicit benefits such as high efficiencies, free maintenances and low noise productions in virtue of the absence of mechanical gears. Prominent features such as high torque density, excellent efficiency and terrific overload capability make permanent-magnet (PM) brushless machine become the cogent candidate for direct drive propulsion.

The permanent-magnet flux-switching (PMFS) machine is a novel double-salient PM brushless machine having both windings and magnets in the stator. The rotor is only a salient passive one and can be robust and fabricated at an easy rate exactly same as a switched reluctance (SR) machine. Consequently, the machine inherits the advantages of both SR machine and PM synchronous machines. Single phase PMFS machines are studied and developed as alternators in airborne application (Rauch and Johnson, 1955), motors for low energy axial fans (Cheng *et al*., 2005) and high speed applications (Chen *et al*., 2006), whilst the most familiar three phase 12/10-pole PMFS machine was first described in (Hoang *et al*., 1997, 2000; Amara *et al*., 2005) further detailed investigations on this machine structure were carried out (Fei and Shen, 2006; Hua and Cheng, 2006; Hua *et al*., 2005, 2006, 2007; Pang, *et al*., 2006, 2007; Zhu, *et al*., 2005, 2007). All the aforementioned literatures have revealed that the machine has some distinct attributes of high torque density, high efficiency, excellent flux-weakening capability and convenience of cooling, which are the exact stringent requirements of EV drive. When compared to conventional PM machine, due to the peculiar locations of the permanent magnets, the PMFS machine exhibits the following advantages: easier to dissipate heat from the stator and therefore, to limit the temperature rise of the magnets; the influence of the armature reaction field on the working point of the magnets is almost negligible since the windings and the magnets are effectively magnetically in parallel. Consequently, the magnets demagnetization hazards owing to over load drive or over temperature are essentially prevented. Additionally, an innovative hybrid excitation flux switching machine based on three phase 12/10-pole structure was presented in (Hoang *et al*., 2007). Moreover, three phase 6/4 and 6/5-pole PMFS machines for high speed operation are investigated in (Fei and Shen, 2006; Wang *et al*., 2007).

Compared with the conventional inner rotor machine, outer rotor machine is intrinsically suitable for direct drive of EV as a result of its low-speed, high-torque features. So far, only the inner rotor PMFS machines are focused in both academic researches and industry applications. The purpose of this paper is to propose and analyze a new outer rotor PMFS machine structure especially for EVs. Firstly, the topology and preliminary sizing equations of the outer rotor PMFS machine are introduced, and the stator and rotor pole number relations defining machine phase configuration are delivered. Secondly, the number of rotor poles are then optimized using comprehensive two-dimensional FEA. Finally, a 12/22-pole three-phase in-wheel outer-rotor PMFS machine is designed, and furthermore, validated and optimized by FEA. The predicted outstanding performance implies that the proposed machine is a leading contender for direct EV drives.

2. Topology

Figure 1 shows the cross-section of a typical 12/22-pole, three-phase outer-rotor PMFS machine. As shown in the figure, the machine is composed of an inner stator that includes stator steel laminations, permanent magnets and armature coils, and an outer salient passive rotor which is exactly same as conventional SR machine, simply constructed by stacked soft magnet steel sheets. Additionally, the concentrated windings, same as SR machine as well, are employed, which result in less copper consumption, high winding fill factor and lower copper Joule loss because of the short end windings. Compared to the conventional outer-rotor PM brushless machines having the magnets in the rotor, the magnets are inset in the middle of the stator poles, which separate the machine stator yoke. Since the magnets and coils are all placed in the stator, the major heat from the machine operation can be easily removed from the stator by various cooling methods, which is desirable for the EV applications where the ambient temperature of the machine may be high. Moreover, the number of pole-pairs in the PMFS machine is the same as the number of rotor teeth; hence, it is easy to achieve a high number of pole-pairs by employing sufficient rotor teeth. This is important for EV propulsion motors which usually require high torque and low speed.

In order to reach sufficient winding area in outer-rotor PMFS machine, the machine main dimension configuration is proposed as $\beta_r=\beta_s=h_{pm}=h_{slot}/5$, as shown in Figure 2, instead of $\beta_r=\beta_s=h_{pm}=h_{slot}$ in conventional inner-rotor PMFS machines (Hoang *et al.*, 1997). Consequently, an adjustment of the main parameter relations to define a polyphased structure must be undertaken and can be demonstrated as follows:

$$N_r = N_s\left(2 \pm \frac{n}{2q}\right) \tag{1}$$

where N_r and N_s are machine rotor and stator pole number respectively, q is number of machine phases and n is a natural number. The relation between the machine mechanical rotation frequency F and the electrical frequency f can be expressed as

$$f = N_r F \tag{2}$$

By virtue of peculiar structure and zero resultant radial stress of the machine, both N_r and N_s should be even numbers. For instance, Figure 1 presents a three phase machine with: q=3; N_s=12; N_r=22.

In PMFS machines, the PM-excited flux always exists and has a constant direction in the magnets. The rotor pole aligns with one of two stator teeth which are embraced by a concentrated winding coil and the PM flux which is linked in the coil goes out of the coil and into the rotor tooth. When the rotor moves forward, the current rotor pole leaves the stator pole and the following rotor pole aligns with the other stator tooth belongs to the same coils, the PM flux linked goes out of the rotor tooth and into the stator tooth. As a result, both magnitudes and polarities of the flux-linkage in the windings will vary periodically along with the rotor moves, which brings the machine outstanding performance.

3. Sizing equations

Analytical sizing equations are usually helpful and necessary during the preliminary stage of machine design, which can significantly improve the machine design efficiency so as to gain the valuable competition time which is exceptionally important in industry. For outer-rotor PMFS machine, the sizing equations can be derived as follows. When the stator outer radius R_{so} is given, the stator tooth width β_s, stator magnet thickness h_{pm}, backiron thickness h_b and slot opening width h_{slot} can be given as follows:

$$\beta_s = h_{pm} = h_b = \frac{h_{slot}}{5} = \frac{\pi R_{so}}{4N_s} \tag{3}$$

Then, the area of one stator slot would be

$$A_s = R_{so}^2 \sin^2\frac{5\pi}{8N_s}\Big/\tan\frac{\pi}{N_s} \tag{4}$$

The electromagnetic torque T_{em} can be obtained as follows:

$$T_{em} = \frac{\frac{\pi}{16} N_r K_d K_p B_g J_{peak} R_{so}^3 l \sin^2\frac{5\pi}{8N_s}}{\tan\frac{\pi}{N_s}} \tag{5}$$

where K_d and K_p are the leakage and winding packing factors respectively, B_g is the peak airgap flux density at no load condition, J_{peak} is the peak current density of the coils, and l is the active length of the machine. The leakage factor in outer-rotor machine is far bigger compared with the one in conventional inner-rotor machine due to its large slot opening. Inspecting equation (5), the machine torque output is proportional to R_{so}^3. R_{so} and l can be analytically gained from equation (5) during the preliminary design stage. In addition, the rotor pole height h_{pr} is chosen as 1/8 stator outer radius R_{so} and rotor yoke thickness h_{yr} is designed as twice the stator back iron thickness h_b for the sake of vibration alleviation. Hence the machine outer radius can be expressed as

$$R_o = \frac{9R_{so}}{8} + g + \frac{\pi R_{so}}{2N_s} \tag{6}$$

where R_o is the rotor outer radius, g is the machine airgap.

4. Optimization of rotor pole numbers

For conventional three-phase inner-rotor PMFS machine with 12 stator poles, the rotor pole number is usually chosen as 10 or 14 to maximize the machine performance. However, since the main dimension configuration has been changed as shown in Figure 2 for outer-rotor PMFS machine, various rotor pole numbers from 14 to 26 for 12 stator pole machine according to equation (1) have been studied to find out the optimal rotor pole number. During the analysis, the rotor outer daimeter, stator dimension, effective axial length and machine operational speed are all kept invariable. The amplitudes of fundamental and 2nd harmonic of back-EMF, as well as peak-to-peak cogging torque with different N_r are listed in Table I. It can be seen that the back-EMF increases along with rotor pole number which is consonant with equation (5), meanwhile significant even harmonics, especially 2nd harmonic, present in the back-EMFs of the machines with 16 and 20 rotor poles due to their asymmetric magnetic structures, additionally, extenssive cogging torque inheres in this two structures. Although it possesses the highest back-EMF and lowest cogging torque, the machine with 26 rotor poles will be operated at the highest electrical frequency that would cause considerable losses. Consequently, the machine with 12 stator poles and 22 rotor poles is considered as the most promising one for the proposed application.

Table I Impacts of rotor pole number on back-EMF and cogging torque

N_r	Fundamental of back-EMF	2nd harmonic of back-EMF	Peak-to-peak Cogging torque
14	1.77 V	0 V	5.23 N·m
16	1.83 V	1.04 V	44.4 N·m
20	2.35 V	0.588 V	13.5 N·m
22	3.06 V	0 V	4.79 N·m
26	3.21 V	0 V	0.945 N·m

5. Machine design and optimization

Machines with 5kW output at rated speed 1000rpm, which are especially suitable for urban vehicle propulsion, have been attracting tremendous attentions from both academics and industry. In this paper, a three phase 12/22-pole structure machine is designed and optimized for certain application. The basic machine dimensions, which is demonstrated in Table II, can be conveniently derived from equations (3) to (6) by substituting T_{em}=50N.m, N_r=22, N_s=10, K_d=0.75, K_p=50%, B_g=2.0T, J_{peak}=7500000A/m^2, l=50mm, and g=0.6mm. Comprehensive FEA are employed to validate the analytical sizing equations, determine the rest machine parameters, and optimize the machine performance. It can be facilely noticed from the FEA results that the back-EMFs of the proposed machine are essentially sinusoidal, which implies that the presented machine is congenitally suitable for BLAC operation. Consequently, the machine performance can be analyzed based on dq-coordinates, and the machine electromagnetic torque can be expressed as

$$T_{em} = \frac{qN_r K_p^2 A_s^2}{8}\left(\phi_{pm} + \left(\Lambda_d - \Lambda_q\right) J_d\right) J_q \tag{7}$$

where ϕ_{pm} is the phase PM flux per turn, Λ_d and Λ_q are the dq- axes permeance per turn, and J_d and J_q are the dq-axes peak current density respectively, which are restricted by

$$J_d^2 + J_q^2 = J_{peak}^2 \tag{8}$$

Machines with various rotor pole width β_r are studied to optimize the machine back-EMF waveform (Hua *et al.*, 2007). In this paper, the rotor pole width is also employed to optimize the machine performance. It should be perceived that the rotor pole width after-mentioned is the normalized value β_r/β_s. Based on FEA, the phase PM flux per turn can be directly derived from open circuit field analysis and the dq-axes permeance per turn can be calculated by the simplified two position method (Hua and Cheng, 2006). From the FEA results as shown in Figure 3, the discrepancy between dq-axes permeance declines while the rotor pole width increases, as a result of machine saliency attenuation. And, the phase PM flux reaches its maximum when the rotor pole width approaches 1.4. The maximum electromagnetic torque and corresponding reluctance component at certain current density can be deduced from equations (7) and (8). Figure 4 shows the variations of electromagnetic and reluctance torque with rotor pole width at rated current density, which are calculated based on the FEA results in

Figure 3. Compared to the total electromagnetic torque, the reluctance component is exiguous even negligible, and meanwhile it diminishes along with the machine rotor pole width escalation and saliency lessening. Furthermore, the electromagnetic torque which the machine could generate at rated current density, resembling the phase PM flux, achieves maximum with the rotor pole width 1.4. Sequentially, the induced voltage (phase back-EMF) obtained for each particular rotor pole width is analyzed and the belt (nontriplen) harmonic distortion (BHD%) in the phase back-EMF is determined. There are only belt harmonics existing in the line-line back-EMF since the triplen harmonics are eradicated internally in three phase machine, and the belt harmonics bring the machine torque ripple which would cause mechanical vibration. The fundamental amplitude of phase back-EMF and BHD% for different rotor pole width are illustrated as Figure 5. Similar to the phase PM flux and electromagnetic torque, the fundamental amplitude of phase back-EMF accomplishes its maximum when the rotor pole width is 1.4. However, BHD% accesses minimum with rotor pole width 1.3. Moreover, Cogging torque, arising from the magnet's tendency to align itself with the minimum reluctance path given by the relative position between rotor and stator, is a parasitic source of mechanical vibration and noise which degrade machine performance, especially at low speed. It also causes startup hesitation for the motor, which is particularly undesirable for traction applications. The peak-to-peak cogging torque variation with different rotor pole width is investigated, as shown in Figure 6. The lowest cogging torque is located at where the rotor pole width is 1.2.

According to the previous analysis, A rotor pole width 1.3 is chosen to achieve the optimal machine performance. So far only the winding turns per coil N_{coil} is still unknown, which can be calculated by the equations in (Hua *et al*., 2006). The PM flux-linkage ψ_{pm} and *dq*-axes inductances L_d/L_q can be derived by

$$\psi_{pm} = N_{coil}\phi_{pm}, \quad L_d = N_{coil}^2\Lambda_d, \quad L_q = N_{coil}^2\Lambda_q \tag{9}$$

Furthermore, the phase resistance can be evaluated by

$$R_{ph} = \rho_{cu}\frac{4N_sN_{coil}^2\left[\left(\frac{5\pi^2}{8N_s}\sin\frac{5\pi}{8N_s}\Big/\sin\frac{\pi}{N_s}+\frac{3\pi}{4N_s}\right)R_{so}+l\right]}{qk_pA_s} \tag{10}$$

The key parameters of the proposed machine are all included in Table II, and the waveforms of phase back-EMF and cogging torque versus rotor mechanical position are depicted as Figure 7.

Table II Machine parameters

Symbol	Machine Parameter	Value	Unit
N_s	Number of Stator Poles	12	
N_r	Number of Rotor Poles	22	
q	Number of Phases	3	
N_{coil}	Number of Winding Turns per Coil	7	
PM	PM Material	NdFeB35	
K_p	Winding Package Factor	50%	
R_{si}	Stator Inner Radius	22.9	mm
R_{so}	Stator Outer Radius	75	mm
R_o	Rotor Outer Radius	94.8	mm
g	Air Gag	0.6	mm
h_{pr}	Rotor Pole Height	9.4	mm
β_s	Stator Tooth Width	3.8	deg
h_{pm}	PM Width	3.8	deg
β_r	Rotor tooth Width	4.9	deg
h_{ps}	Stator Tooth Width	47.2	mm
l	Machine Active Axial Length	50	mm
J_{peak}	Rated Peak Current Density	7500000	A/m^2
U_{dc}	DC-link Voltage	42	V

n	Rated Speed	1000	rpm
P_{em}	Rated Machine Power output	5.2	kW
I_{peak}	Rated Peak Phase Current	152	A
Ψ_{pm}	Phase PM Flux Linkage	10.2	mWb
L_d	d-axis Inductance	62.7	μH
L_q	q-axis Inductance	72.0	μH
R_{ph}	Phase Resistance (100 ^{o}C)	11.1	mΩ

6. Machine losses analysis

Machine losses are a complex function of speed and load. Electromagnetic losses dominate total losses in low speed machine, consequently, this paper is only concentrated on the analysis of electromagnetic losses. Electromagnetic losses can be broken down into three distinct parts, copper losses in the machine coils, core losses in the stator and rotor laminations, and eddy current losses in the permanent magnets. The motor coil copper losses can be computed from the estimated phase resistance and the torque-current profile of the machine from the FEA static analysis. The estimated resistance of one phase of the machine at 100 ^{o}C is 11.1 mΩ. Using the predicted current densities required to achieve different torque values, the copper losses for four load conditions are given in Table III.

Table III Copper losses

Torque	Phase Peak Current	Loss
25 N.m	76 A	96 W
50 N.m	152 A	383 W
75 N.m	228 A	862 W

Transient FEA can be used to calculate the core losses in electrical steel laminations considering the harmonics. At a given frequency, the core losses for electrical steel is

$$P_c = K_h B_{\max}^2 f + K_c \left(B_{\max} f\right)^2 + K_e \left(B_{\max} f\right)^{1.5} \tag{11}$$

where K_h, K_c, and K_e are hysteresis loss, classic eddy current loss, and excess or anomalous eddy current loss coefficients, respectively, which are all information of the material from manufacturer. B_{max} is the maximum amplitude of flux density. In order to reduce the core losses in the machine, fairly thin laminations (0.2mm) are employed. Figure 8 shows the sensitivity of the lamination core losses to the machine speed and load. Increasing the load of the machine, the armature current increases the peak flux density in the lamination which leads to higher core losses. The peak load condition will have approximately double the core losses compared to the no load condition. However, the core losses will not be significant compared with copper losses and eddy current losses in PM.

The magnetic field in PM will vary when the rotor rotates, since NdFeB35 has conductivity of roughly 10 times less than mild steel, which can generate considerable eddy current losses. Figure 9 shows that the eddy current losses in the PM of the proposed machine are significant, which are nearly two times the copper losses under the rated load condition. It can be envisaged that the machine with 26 rotor poles would inhere in even much larger eddy current losses in permanent magnets to deteriorate the machine performance further. Reducing eddy current losses in PM is important, not only because the magnets have a low maximum working temperature, but also because the energy product of the magnets is notably condensed as their temperature is increased. So researches on the PM eddy current losses as well as the means to reduce the PM eddy current losses in outer-rotor PMFS machine is especially necessary.

From the electromagnetic loss data presented in this section, an efficiency map has been compiled, as shown in Figure 10. It can be found the efficiency of the machine under rated load condition is nearly 84%. The copper losses in winding coils and eddy current losses in PM are the major components of electromagnetic losses, which should be reduced to improve the machine efficiency. It is worth mentioning that the eddy current losses in PM grow to be the biggest part along as the operating speed increases. The eddy current losses can be effectively shrunk by segmenting the magnets into several sections (Zhu, *et al*., 2008), additionally the copper losses can be allayed by improve the machine winding packing factor. By the means aforementioned, the machine efficiency can overtake 90% to further confirm the proposed outer-rotor PMFS machine as a potential contender for EV propulsion.

7. Machine losses analysis

A novel outer-rotor PMFS machine is proposed for electric vehicle propulsion in this paper. The machine topology is introduced first, and the machine sizing equations are developed for preliminary design, which is validated by FEA. A 5kW machine with optimal 22 rotor poles is designed based on the analytical model and moreover, FEA is employed to optimize the machine performance and predict the electromagnetic losses. It is concluded that the machine possesses several distinct advantages that underpin the machine as a leading candidate for EV application, even though the machine efficiency is not as good as desired. Eddy current losses in PM minimization techniques will be investigated in the future works, as well as the reduction methods of copper losses in winding coils.

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