

Article

Analysis of a Five-Phase Fault-Tolerant Consequent-Pole Permanent-Magnet Vernier Machine

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Abstract

The five-phase fault-tolerant consequent-pole permanent-magnet Vernier machine (FTCP-PMVM) has attracted extensive research attention owing to its excellent permanent-magnet utilization while maintaining competitive electromagnetic performance. However, the double-salient structure poses considerable challenges for analysis. The torque generation and power factor characteristics of the FTCP-PMVM are analyzed from a magnetic field modulation perspective in this work. Initially, based on the air gap field modulation effect, the modulation processes of both the permanent-magnet field and the armature field are analyzed. Subsequently, the torque generation mechanism is explained through harmonic matching resulting from the field modulation process, the results demonstrate that the 23rd air gap harmonic dominates the generation of average electromagnetic torque and accounts for the majority of output torque. Furthermore, the power factor is examined in depth by analyzing the reactive power contributed by the machine inductive components (e.g., self-inductance, mutual inductance and leakage inductance) and the active power generated by the permanent-magnet portion, all from the standpoint of field modulation. Finally, a prototype is fabricated to test the machine's torque, power factor and efficiency. Experimental data confirms the reliability of the theoretical analysis.

Keywords: consequent-pole; fault-tolerant; permanent-magnet Vernier machine; magnetic field modulation effect; mechanism of torque generation; inductance; power factor

1. Introduction

Driven by rising demands for high-efficiency, high-torque-density electrical machines, permanent-magnet synchronous machines have been widely adopted in low-speed high-torque applications including electric vehicles, marine propulsion systems, wind turbine generators and renewable energy generators [1–4]. However, due to the magnetic field modulation effect, permanent-magnet Vernier machines have gained increasing attention for their higher torque density and more compact size [5,6].

Due to the scarcity of the PMs and the inter-magnet flux leakage, in [7], Li et al. proposed that consequent-pole permanent-magnet Vernier machines (CP-PMVMs) exhibit higher main flux and lower leakage flux compared with conventional permanent-magnet Vernier machines. Moreover, the topology enables higher torque density and back-EMF with decreased permanent-magnet consumption. Such a double-salient topology not only cuts permanent-magnet consumption by half but also effectively restrains inter-magnet leakage flux, thereby further boosting torque density and magnet utilization efficiency [8]. The contribution of CP-PMVM is analyzed in [9], where the modulated field harmonics



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in CP-PMVM contribute higher torque proportion than their SM-PMVM ones, which indicates that the CP-PMVM exhibit enhanced magnetic field modulation effect due to its double-side slotting effect and unipolar PM MMF distributions. To improve torque density, a consequent-pole V-shape permanent-magnet Vernier machine is proposed [10]. An alternating flux barrier structure is placed on the bottom of pair-wise spoke-type magnets, which can improve the flux path and create an additional rotor magnetic circuit. The results show that the back-EMF becomes more sinusoidal, and the torque is also increased [11]. Dual-PM Vernier machines with two pieces on each stator tooth are proposed. By comparing four different topologies, the results indicate that each machine is characterized by distinct dominant harmonics. Among them, the N/Fe/N/Fe arrangement exhibits the highest electromagnetic torque and the superior efficiency [12]. Hybrid CP-PMVM is proposed to improve the overload performance by adding SPMs into the ICPM rotor [13]. A consequent-pole dual-rotor permanent-magnet Vernier machine could produce a higher power factor and torque density [14].

On the other hand, in high-reliability direct-drive applications such as aerospace, marine propulsion and autonomous driving, the fault-tolerant capability of electric machines has emerged as a critical metric for evaluating system performance. Currently, effective strategies to enhance the fault tolerance of permanent-magnet machines primarily include implementing physical and magnetic isolation to achieve low mutual inductance, thereby decoupling faulty phases from healthy ones [15,16]; increasing winding turns or optimizing slot tooth geometries to realize high self-inductance [17], which effectively suppresses single-phase short-circuit currents; and employing multi-phase winding configurations to increase system operational redundancy [18–20]. Zhao et al. provide a holistic approach to analyze and design FT-PMVM. The electromagnetic performance of the fault-tolerant machine is thoroughly analyzed by considering the field modulation effect and the equivalent magnetic circuit model [21].

Existing literature separately investigates CP-PMVM and fault-tolerant PMVM topologies but only evaluates isolated performance metrics. Few studies systematically quantify the coupling between air gap field harmonics, electromagnetic torque and power factor, nor clarify the inherent linkage between inductance parameters and power factor behavior. To maintain high permanent-magnet utilization while fulfilling the stringent requirements for fault suppression in mission-critical applications, this paper investigates a five-phase fault-tolerant consequent-pole permanent-magnet Vernier machine (FTCP-PMVM). Despite the prominent advantages of FTCP-PMVM in high-performance direct-drive applications, their intrinsic double-salient structure excites highly complex harmonic components within the air gap magnetic field. Consequently, a comprehensive analysis of the torque generation mechanism and power factor based on the magnetic field modulation effect remains largely unexplored, making the precise analysis of torque generation mechanisms and power factor highly challenging. Thus, research on the mechanisms of torque and power factor generation holds significant importance.

The remainder of this paper is organized as follows. Section 2 introduces the modulation process of both the permanent-magnet field and the armature field with the double-salient structure, analyzing the harmonics and their respective speeds generated by these two fields. Based on the harmonic orders and speeds analyzed in Section 2, Section 3 elaborates on the mechanism of torque generation in detail, evaluating the contribution of each harmonic component. Section 4 analyzes the power factor of the FTCP-PMVM. By researching the reactive power arising from machine inductances, such as self-inductance, mutual inductance and leakage inductance, as well as the active power generated by the permanent-magnet portion, this section details the analytical process for determining the power factor.

2. Topology and Field Modulation Mechanism of FTCP-PMVM

Figure 1 illustrates the topology of a 20-slot/23-pole fault-tolerant consequent-pole permanent-magnet Vernier machine (FTCP-PMVM). The rotor employs a consequent-pole configuration consisting of unipolarly magnetized PMs and salient iron poles, where the iron poles are magnetized to function as equivalent poles of opposite polarity, thereby forming complete pole pairs. On the stator side, a structure featuring alternating fault-tolerant teeth and split teeth is adopted to enhance the air gap magnetic field modulation effect. To enhance the structural compactness, five-phase concentrated windings are wound exclusively around the split teeth, while the fault-tolerant teeth serve as inter-phase isolation without windings. Five different colors are used to represent the currents of five phases respectively. The five-phase FTCP-PMVM can operate continuously and stably under multi-phase open-circuit faults, featuring low torque ripple and an excellent balance between fault-tolerant capability and control complexity. Table 1 gives the main design parameters of the machine.

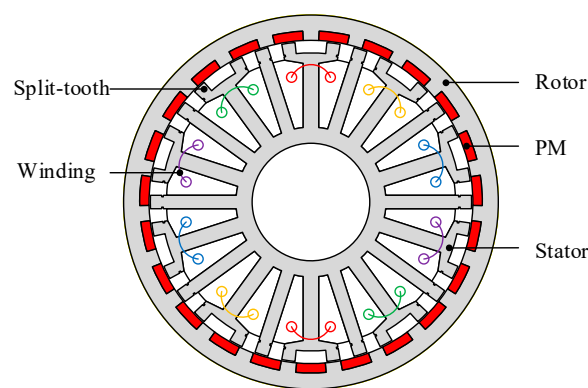


Figure 1. Topology structure of the FTCP-PMVM.

Table 1. The Parameters of FTCP-PMVM.

Items	Units	Value
Rotor outer radius	mm	70
Stator inner radius	mm	22
Stator outer radius	mm	60
Axial length	mm	60
Air gap length	mm	0.5
Rated current	A	10.6
rated speed	r/min	600
Winding turns per phase	/	45
Permanent magnet material type	/	NdFe35
Silicon steel sheet grade	/	DW465_50

2.1. Modeling of PM Field

For modeling convenience, the following assumptions are made:

1. The magnetomotive force produced by permanent magnets is equivalent to an ideal square wave.
2. The stator and rotor iron cores operate in the linear region, and a linear permeance modulation model is adopted.
3. Two-dimensional electromagnetic field analysis is adopted in this study for simplicity, and winding end leakage effects are neglected in the MMF modeling process.

Figure 2 illustrates the modulation of PM MMF of the FTCP-PMVM. The initial PM MMF distribution is assumed to be an ideal square wave, as shown in the green wave. The modulation process is divided into two steps: the initial PM MMF is modulated by the stator permeance, and subsequently, it is modulated by the rotor permeance.

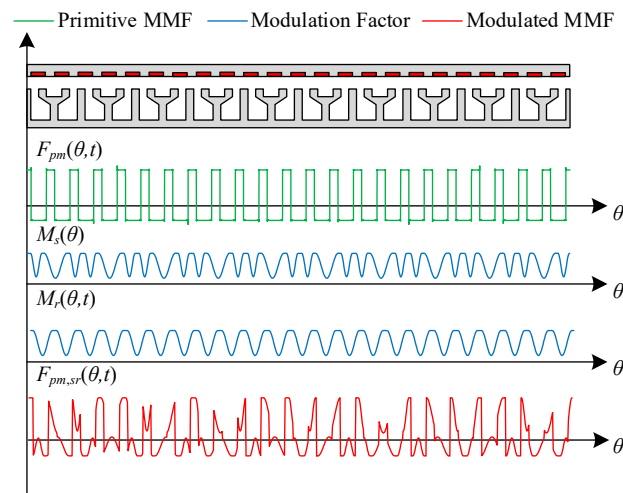


Figure 2. The process of PM MMF modulation of FTCP-PMVM.

The rotor varies with both time and mechanical angle in the magnetic field, which can be expressed by Fourier series as

$$M_r(\theta, t) = a_{r,0} - \sum_{i=1}^{\infty} a_{r,i} \cos[iP_r(\theta - \Omega_r t)] \quad (1)$$

where $a_{r,0}$ and $a_{r,i}$ are the zero-order and iP_r -order harmonic amplitudes of rotor permeance. P_r is PM pole pair. θ and Ω_r represent mechanical angle and mechanical angular velocity, respectively. i is the positive integer.

Since the stator is stationary in the magnetic field, and the stator only depends on the mechanical angle, which can be expressed by Fourier series as

$$M_s(\theta) = a_{s,0} - \sum_{j=1}^{\infty} a_{s,j} \cos(jP_s\theta) \quad (2)$$

where $a_{s,0}$ and $a_{s,j}$ are the zero-order and jP_s -order harmonic amplitudes of stator permeance. For the hybrid stator structure of the FTCP-PMVM, the stator tooth unit consists of a main tooth containing two split teeth and a fault-tolerant tooth. Thus, the number of stator tooth unit P_s is equal to 10. j is the positive integer.

The Fourier series of the consequent-pole PM MMF can be expressed as

$$F_{pm}(\theta, t) = \sum_{k=1}^{\infty} F_{pm,k} \cos[kP_r(\theta - \Omega_r t)] \quad (3)$$

where $F_{pm,k}$ is the amplitude of k -order consequent PM MMF harmonic.

By the modulation of stator and rotor modulation factor, the PM air gap flux density B_{pm} can be calculated by the product of (3) and (1) and (2), and it yields

$$\begin{aligned}
 B_{pm}(\theta, t) &= \frac{\mu_0}{g} F_{pm}(\theta, t) M_s(\theta) M_r(\theta, t) \\
 &= \underbrace{\frac{\mu_0}{g} a_{s,0} a_{r,0} \sum_{k=1}^{\infty} F_{pm,k} \cos[kP_r(\theta - \Omega_r t)]}_{\text{Unit modulation}} \\
 &\quad - \underbrace{\frac{\mu_0}{2g} a_{r,0} \sum_{k=1}^{\infty} \sum_{j=1}^{\infty} F_{pm,k} a_{s,j} \cos[(kP_r \pm jP_s)\theta - kP_r \Omega_r t]}_{\text{Stator modulation}} \\
 &\quad - \underbrace{\frac{\mu_0}{2g} a_{s,0} \sum_{k=1}^{\infty} \sum_{i=1}^{\infty} F_{pm,k} a_{r,i} \cos[(kP_r \pm iP_r)(\theta - \Omega_r t)]}_{\text{Rotor modulation}} \\
 &\quad + \underbrace{\frac{\mu_0}{4g} \sum_{k=1}^{\infty} \sum_{i=1}^{\infty} \sum_{j=1}^{\infty} F_{pm,k} a_{r,i} a_{s,j} \cos[(kP_r + iP_r \pm jP_s)\theta - (kP_r + iP_r)\Omega_r t]}_{\text{Rotor and stator modulation}} \\
 &\quad + \frac{\mu_0}{4g} \sum_{k=1}^{\infty} \sum_{i=1}^{\infty} \sum_{j=1}^{\infty} F_{pm,k} a_{r,i} a_{s,j} \cos[(kP_r - iP_r \pm jP_s)\theta - (kP_r - iP_r)\Omega_r t]
 \end{aligned} \tag{4}$$

where μ_0 is the vacuum permeability and g is the length of air gap.

To validate the modeling accuracy, it is necessary to compare the analytical air gap flux density with the finite element results. As shown in Figure 3, the finite element results exhibit good agreement with the analytical results, demonstrating the effectiveness of the model establishment.

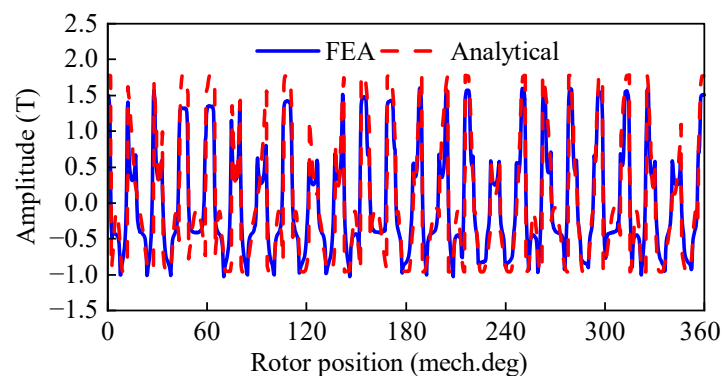


Figure 3. The waveform of PM air gap flux density of FTCP-PMVM.

Figure 4 shows the dominant harmonics in PM flux density. It can be seen that the amplitude of 3rd, 7th, 23rd and 27th harmonics are significant. It can be seen that the harmonic orders are quite abundant and the finite element analysis shows high consistency with the analytical results.

Table 2 lists the harmonics and their speed that appear during the modulation process of PM MMF. It can be observed that through the dual modulation effects of the stator and rotor, the PM field generates various harmonic components, characterized by the orders kP_r , $|kP_r \pm jP_s|$, $|kP_r \pm iP_r|$ and $|kP_r \pm iP_r \pm jP_s|$. For instance, considering the 3rd, 7th, 23rd, 27th harmonics, their respective rotational velocities are identified as $23/3\Omega_r$, $-23/7\Omega_r$, Ω_r , $-23/17\Omega_r$.

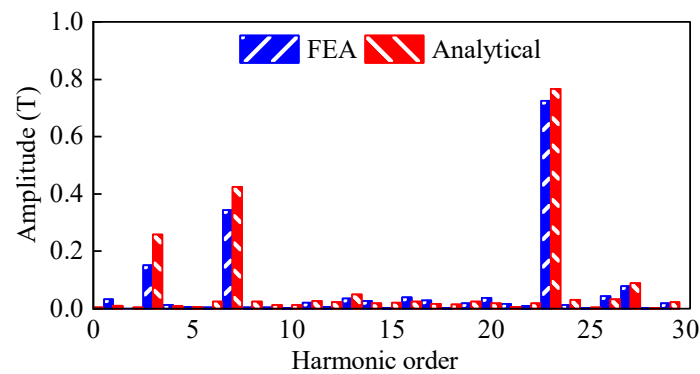


Figure 4. Harmonics spectra of PM air gap flux density of FTCP-PMVM.

Table 2. The Distribution and Speed of Harmonic in PM Field.

Item	Order	Speed	Examples	
			Order	Speed (Ω_r)
PM field	kP_r	Ω_r	23rd, 46th	1
	$kP_r + jP_s$	$\frac{kP_r\Omega_r}{kP_r + jP_s}$	33rd, 43rd	23/33, 23/43
	$ kP_r - jP_s $	$\frac{kP_r\Omega_r}{kP_r - jP_s}$	13th, 3rd, 7th, 17th	23/13, 23/3, -23/7, -23/17
	$kP_r + iP_r$	Ω_r	46th	1
	$ kP_r - iP_r $	Ω_r	23rd	1
	$kP_r + iP_r + jP_s$	$\frac{(kP_r + iP_r)\Omega_r}{kP_r + iP_r + jP_s}$	56th, 66th	46/56, 46/66
	$ kP_r + iP_r - jP_s $	$\frac{(kP_r + iP_r)\Omega_r}{kP_r + iP_r - jP_s}$	36th, 26th	46/36, 46/26
	$ kP_r - iP_r + jP_s $	$\frac{(kP_r - iP_r)\Omega_r}{kP_r - iP_r + jP_s}$	10th, 20th, 30th	0, 0, 0
	$ kP_r + iP_r - jP_s $	$\frac{(kP_r - iP_r)\Omega_r}{kP_r + iP_r - jP_s}$	13th, 3rd	23/13, 23/3

Figure 5 shows the dominant harmonics and their rotational directions in the radial PM flux density, including the 3rd, 7th and 23rd order harmonics. It can be seen that, assuming the direction of the 23rd harmonic is positive, the 3rd harmonic rotates in the same direction as the 23rd harmonic, while the 7th harmonic rotates in the opposite direction. This is consistent with the analytical predictions in Table 2.

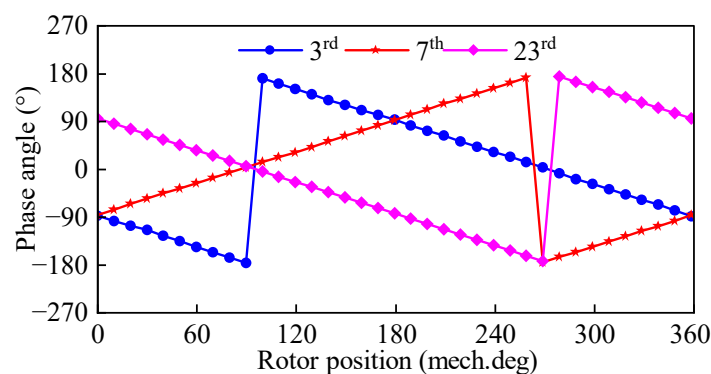


Figure 5. Phase of PM air gap flux density of FTCP-PMVM.

2.2. Modeling of Armature Field

The machine is excited by five-phase windings, and the winding function can be expressed by Fourier series as

$$\begin{cases} N_A(\theta) = \sum_{v=1}^{\infty} N_v \cos(vP_w\theta) \\ N_B(\theta) = \sum_{v=1}^{\infty} N_v \cos[vP_w(\theta - 4\pi/5)] \\ N_C(\theta) = \sum_{v=1}^{\infty} N_v \cos[vP_w(\theta + 2\pi/5)] \\ N_D(\theta) = \sum_{v=1}^{\infty} N_v \cos[vP_w(\theta - 2\pi/5)] \\ N_E(\theta) = \sum_{v=1}^{\infty} N_v \cos[vP_w(\theta + 4\pi/5)] \end{cases} \quad (5)$$

where N_v is the winding function, P_w is the pole pair number of windings.

The current is excited by five-phase sinusoidal current, which can be expressed by

$$\begin{cases} I_A(t) = I_{rms} \sin(\omega t) \\ I_B(t) = I_{rms} \sin(\omega t - 2\pi/5) \\ I_C(t) = I_{rms} \sin(\omega t - 4\pi/5) \\ I_D(t) = I_{rms} \sin(\omega t + 4\pi/5) \\ I_E(t) = I_{rms} \sin(\omega t + 2\pi/5) \end{cases} \quad (6)$$

where I_{rms} is the root mean square value of the current.

Similar to the modulation of PM field, the modulation process of armature field is also divided into two steps, as shown in Figure 6: the initial armature MMF is first modulated by the stator permeance to yield $F_{am,s}$ and then modulated by the rotor permeance to produce $F_{am,sr}$.

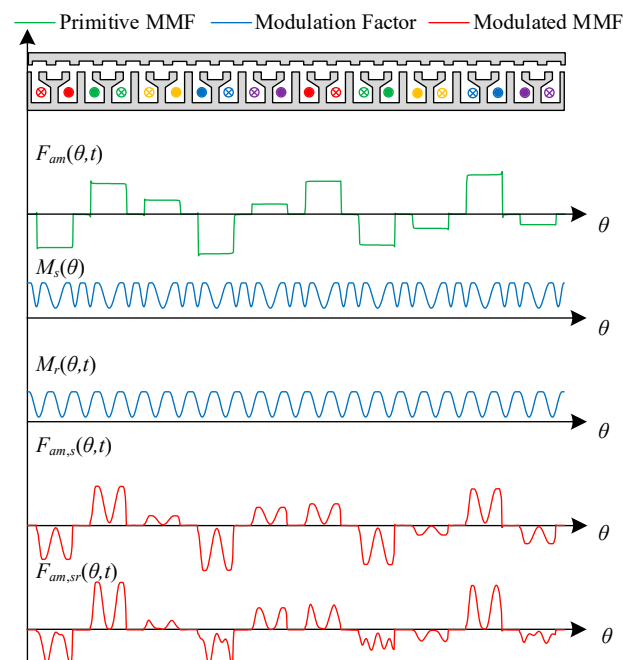


Figure 6. The process of armature MMF modulation of FTCP-PMVM.

The initial armature MMF is generated by the product of (5) and (6), which can be expressed as

$$\begin{aligned}
 F_{aw} &= N_A(\theta)I_A(t) + N_B(\theta)I_B(t) + N_C(\theta)I_C(t) \\
 &+ N_D(\theta)I_D(t) + N_E(\theta)I_E(t) \\
 &= \frac{5}{2} \sum_{v=7,17,27,\dots}^{\infty} N_v I_{rms} \sin(vP_w\theta + \omega t) - \frac{5}{2} \sum_{v=3,13,23,\dots}^{\infty} N_v I_{rms} \sin(vP_w\theta - \omega t) \\
 &= \frac{5}{2} \sum_{v=10l-3}^{\infty} N_v I_{rms} \sin(vP_w\theta + \omega t) - \frac{5}{2} \sum_{v=10l-7}^{\infty} N_v I_{rms} \sin(vP_w\theta - \omega t) \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots \\
 &= \sum_{v=10l-3}^{\infty} F_v \sin(vP_w\theta + \omega t) - \sum_{v=10l-7}^{\infty} F_v \sin(vP_w\theta - \omega t) \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots
 \end{aligned} \tag{7}$$

where F_v is the vP_w -order space harmonic amplitude of armature winding MMF.

Then, similar to the generation process of the PM air gap flux density, the armature air gap flux density B_{aw} can be derived by multiplying armature MMF and both stator and rotor modulation factor, which can be written as

$$\begin{aligned}
 B_{aw}(\theta, t) &= \frac{\mu_0}{g} F_{aw}(\theta, t) M_s(\theta) M_r(\theta, t) \\
 &= \frac{\mu_0}{g} a_{s,0} a_{r,0} \sum_{v=10l-3}^{\infty} F_v \sin(vP_w\theta + P_r\Omega_r t) - \frac{\mu_0}{g} a_{s,0} a_{r,0} \sum_{v=10l-7}^{\infty} F_v \sin(vP_w\theta - P_r\Omega_r t) \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots \\
 &\quad \underbrace{\hspace{15em}}_{\text{Unit modulation}} \\
 &- \frac{\mu_0}{2g} a_{r,0} \sum_{v=10l-3}^{\infty} \sum_{i=1}^{\infty} F_v a_{s,j} \sin[(vP_w \pm jP_s)\theta + P_r\Omega_r t] + \frac{\mu_0}{2g} a_{r,0} \sum_{v=10l-7}^{\infty} \sum_{i=1}^{\infty} F_v a_{s,j} \sin[(vP_w \pm jP_s)\theta - P_r\Omega_r t] \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots \\
 &\quad \underbrace{\hspace{15em}}_{\text{Stator modulation}} \\
 &- \frac{\mu_0}{2g} a_{s,0} \sum_{v=10l-3}^{\infty} \sum_{i=1}^{\infty} F_v a_{r,i} \sin[(vP_w \pm iP_r)\theta + (1 \pm i)P_r\Omega_r t] + \frac{\mu_0}{2g} a_{s,0} \sum_{v=10l-7}^{\infty} \sum_{i=1}^{\infty} F_v a_{r,i} \sin[(vP_w \pm iP_r)\theta - (1 \pm i)P_r\Omega_r t] \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots \\
 &\quad \underbrace{\hspace{15em}}_{\text{Rotor modulation}} \\
 &+ \frac{\mu_0}{4g} \sum_{v=10l-3}^{\infty} \sum_{i=1}^{\infty} \sum_{j=1}^{\infty} F_v a_{s,j} a_{r,i} [(vP_w \pm jP_s \pm iP_r)\theta - (1 \pm i)P_r\Omega_r t] - \frac{\mu_0}{4g} \sum_{v=10l-7}^{\infty} \sum_{i=1}^{\infty} \sum_{j=1}^{\infty} F_v a_{s,j} a_{r,i} [(vP_w \pm jP_s \pm iP_r)\theta + (1 \pm i)P_r\Omega_r t] \\
 &\quad l = 1, 2, 3 \dots \quad l = 1, 2, 3 \dots \\
 &\quad \underbrace{\hspace{15em}}_{\text{Stator and rotor modulation}}
 \end{aligned} \tag{8}$$

Figure 7 illustrates the finite element analysis and analytical results of armature flux density. The consistency between the finite element analysis and the analytical results validates the accuracy of the proposed model.

Figure 8 shows the distribution of harmonics during the modulation process of armature MMF. The 3rd, 7th, 17th, 20th, 23rd and 27th harmonics exhibit significant amplitudes and play dominant role in the armature modulation process. Meanwhile, the analytical results show good agreement with the finite element analysis result.

Table 3 lists the harmonics and their rotational speed that arise during the modulation process of armature MMF.

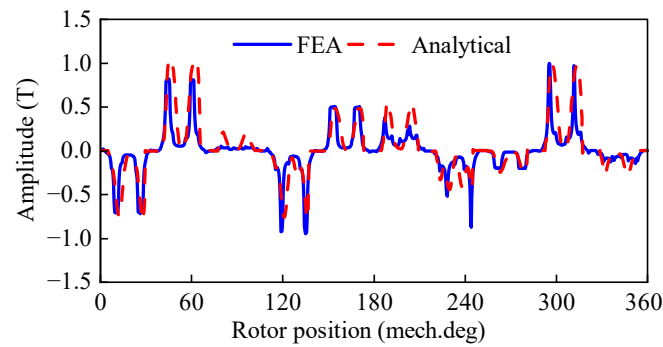


Figure 7. The waveform of armature flux density of FTCP-PMVM.

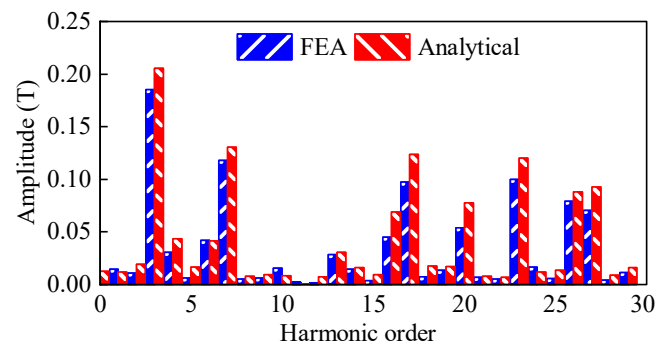


Figure 8. Harmonics spectra of armature flux density of FTCP-PMVM.

Table 3. The Distribution and Speed of Harmonic in Armature Field.

Item	Order	Speed	Examples		Item	Order	Speed	Examples	
			Order	Speed (Ω_r)				Order	Speed (Ω_r)
$v = 10l - 3$ $l = 1, 2, 3, \dots$	vP_w	$\frac{-P_r\Omega_r}{vP_w}$	7th, 17th	$-\frac{23}{7}, -\frac{23}{17}$	$v = 10l - 7$ $l = 1, 2, 3, \dots$	vP_w	$\frac{P_r\Omega_r}{vP_w}$	3rd, 13th	$\frac{23}{3}, \frac{23}{13}$
	$vP_w + jP_s$	$\frac{-P_r\Omega_r}{vP_w + jP_s}$	17th, 27th	$-\frac{23}{17}, -\frac{23}{27}$		$vP_w + jP_s$	$\frac{P_r\Omega_r}{vP_w + jP_s}$	13th, 23rd	$\frac{23}{13}, 1$
	$ vP_w - jP_s $	$\frac{-P_r\Omega_r}{vP_w - jP_s}$	3rd, 13th	$\frac{23}{3}, \frac{23}{13}$		$ vP_w - jP_s $	$\frac{P_r\Omega_r}{vP_w - jP_s}$	7th, 17th	$-\frac{23}{7}, -\frac{23}{17}$
	$vP_w + iP_r$	$\frac{-(1+i)P_r\Omega_r}{vP_w + iP_r}$	30th, 40th	$-\frac{46}{30}, -\frac{46}{40}$		$vP_w + iP_r$	$\frac{(1+i)P_r\Omega_r}{vP_w + iP_r}$	26th	$\frac{46}{26}$
	$ vP_w - iP_r $	$\frac{-(1-i)P_r\Omega_r}{vP_w - iP_r}$	16th, 6th	0, 0		$ vP_w - iP_r $	$\frac{(1-i)P_r\Omega_r}{vP_w - iP_r}$	20th, 10th	0, 0
	$vP_w + jP_s + iP_r$	$\frac{-(1-i)P_r\Omega_r}{vP_w + jP_s + iP_r}$	40th	0		$vP_w + jP_s + iP_r$	$\frac{(1-i)P_r\Omega_r}{vP_w + jP_s + iP_r}$	36th	0
	$ vP_w + jP_s - iP_r $	$\frac{-(1+i)P_r\Omega_r}{vP_w + jP_s - iP_r}$	4th, 14th	$-\frac{46}{4}, -\frac{46}{14}$		$ vP_w + jP_s - iP_r $	$\frac{(1+i)P_r\Omega_r}{vP_w + jP_s - iP_r}$	10th	$\frac{46}{10}$
	$ vP_w - jP_s + iP_r $	$\frac{-(1-i)P_r\Omega_r}{vP_w - jP_s + iP_r}$	20th, 30th	0, 0		$ vP_w - jP_s + iP_r $	$\frac{(1-i)P_r\Omega_r}{vP_w - jP_s + iP_r}$	16th	0
	$ vP_w - jP_s - iP_r $	$\frac{-(1+i)P_r\Omega_r}{vP_w - jP_s - iP_r}$	26th	$\frac{46}{26}$		$ vP_w - jP_s - iP_r $	$\frac{(1+i)P_r\Omega_r}{vP_w - jP_s - iP_r}$	30th, 20th	$-\frac{46}{30}, -\frac{46}{20}$

Similar to the PM magnetic field, the armature magnetic field also exhibits four types of modulated harmonics, vP_w , $|vP_w \pm jP_s|$, $|vP_w \pm iP_r|$ and $|vP_w \pm jP_s \pm iP_r|$. It can be seen that abundant harmonics are modulated, such as 3rd, 7th, 10th, 13th, 17th, 20th and 23rd, with speeds of $\frac{23}{3}\Omega_r$, $-\frac{23}{7}\Omega_r$, 0, $\frac{23}{13}\Omega_r$ and $-\frac{23}{17}\Omega_r$, 0, Ω_r .

3. Quantitative Analysis of Air Gap Field Harmonics to Torque Contribution

3.1. Torque Generation Mechanism

PM flux linkage is the product of the MMF of PM and the winding function, which can be deduced by

$$\psi_{pm}(t) = \frac{\mu_0 r_g l_{st}}{g} \int_0^{2\pi} F_{pm, sr}(\theta, t) \cdot N(\theta) d\theta \quad (9)$$

where the r_g is the air gap radius and l_{st} denotes the stack length.

Figure 9 demonstrates the agreement between the analytical results and the finite element analysis of PM flux linkage waveform. It can be observed that the amplitude of PM flux linkage is 0.05 Wb, and the error between them is very small.

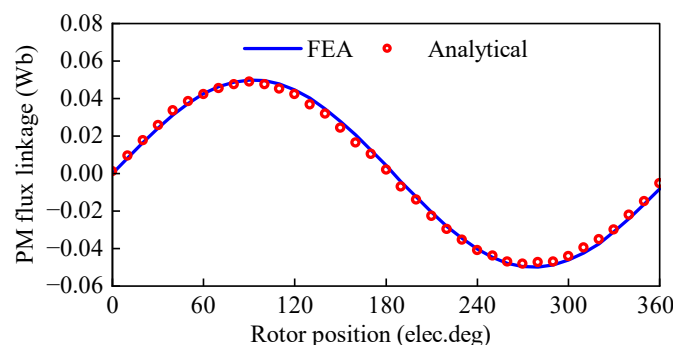


Figure 9. Waveform of PM flux linkage of FTCP-PMVM.

The PM torque can be written as

$$T_{pm}(t) = P_r \sum_{n=A,B,C,D,E} \psi_{pm,n}(t) I_n(t) \quad (10)$$

Figure 10 shows the waveform of torque of finite element analysis and analytical results. The average torque of finite element analysis is 26.36 Nm, while the average torque of analytical result yields an average torque of 26.23. The error between them is within 5%, which validates the effectiveness of the model establishment.

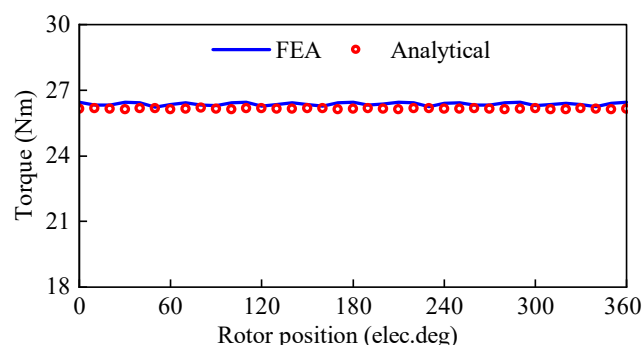


Figure 10. Waveform of torque of FTCP-PMVM.

According to classical electromagnetic theory, torque is generated only when the harmonics produced by the armature field and PM field have the same order and rotate at the same speed. Table 4 lists the identical harmonics and speed produced by both the armature field and PM field. It can be observed that the harmonic contributing to the electromagnetic torque is the 3rd, 7th, 10th, 13th, 17th, 23rd, 27th, 30th, 43rd, 46th and 47th. Since the amplitudes of harmonics above the 50th order are relatively low, their contribution to electromagnetic torque is negligible and can be reasonably ignored.

Table 4. Partial harmonics with identical order and speed.

Order	3	7	10	13	17	23	27	30	43	46	47
Speed	23/3	−23/7	0	23/13	−23/17	1	−23/27	0	23/43	1	−23/47

3.2. Calculation of Torque Contribution

In order to calculate the contribution of each harmonic component to the electromagnetic torque, the maxwell stress tensor method can be employed. Based on maxwell stress tensor method, the electromagnetic torque of rotating machine can be calculated as

$$T = \frac{r_g^2 l_{st}}{u_0} \int_0^{2\pi} B_r B_t d\theta \quad (11)$$

where r_g and l_{st} represent air gap radius and stack length, respectively. B_r and B_t denote the radial and tangential components of the air gap magnetic flux density, respectively, which can be expressed in terms of Fourier coefficients as follows

$$\begin{cases} B_r = \sum_m B_{rm} \cos[m\theta - \theta_{rm}(t)] \\ B_t = \sum_n B_{tn} \cos[n\theta - \theta_{tn}(t)] \end{cases} \quad (12)$$

where B_{rm} and B_{tn} are the amplitudes of m th and n th of radial and tangential air gap flux density. θ_{rm} and θ_{tn} are the corresponding phase angles.

Substituting (12) into (11), it yields

$$T = \frac{\pi r_g^2 l_{st}}{u_0} \sum_{m=n} B_{rm} B_{tn} \cos[\theta_{rm}(t) - \theta_{tn}(t)] \quad (13)$$

By substituting the amplitude and phase of each air gap flux density harmonics into (13), the electromagnetic torque contributions can be obtained, as shown in Figure 11. It can be observed that the 23rd harmonic dominates the average torque generation.

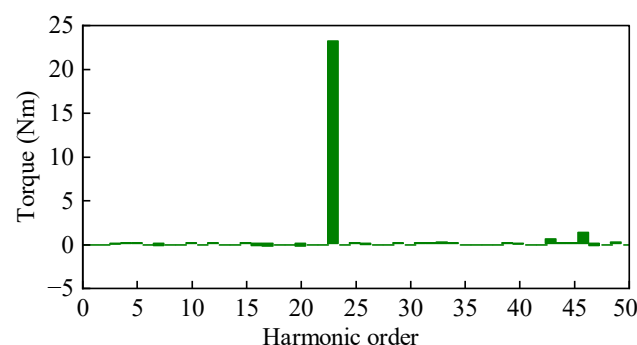
**Figure 11.** Proportion of harmonics torque contribution.

Figure 12 illustrates the contribution of each harmonic to the torque over one electrical cycle. As can be seen, similar to Figure 11, the 23rd harmonic order provides a substantial and steady torque contribution throughout the entire period.

Table 5 lists the harmonic orders, rotational speeds and their contribution of each harmonic to the total torque. The modulation sources of these harmonics originating from both the armature magnetic field and the permanent-magnet magnetic field are also indicated. It can be seen that the 10th and 23rd harmonics generate positive torque contributions, whereas all other harmonics yield negative torque components.

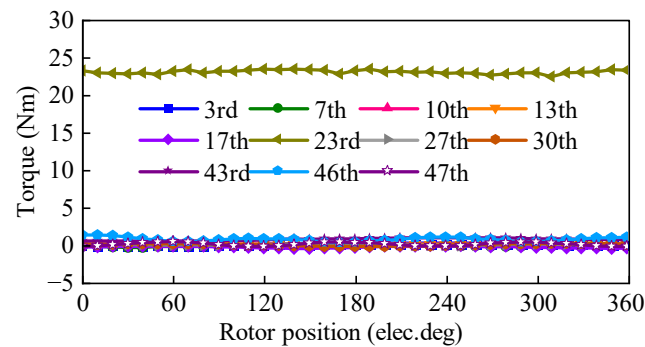


Figure 12. The waveform of torque contribution over one electrical cycle.

Table 5. The proportion of harmonics contribution and their source.

Order	PM Field	Armature Field	Speed (Ω_r)	Contribution (Nm)	Proportion (%)
7	$ P_r - 3P_s $	$7P_w$	$-23/7$	-0.26	-0.97
17	$ P_r - 4P_s $	$17P_w/7P_w + 10P_s/27P_w - 10P_s$	$-23/17$	-0.33	-1.25
23	$P_r/ 2P_r - P_r $	$ 33P_w - P_s /23P_w/13P_w + P_s$	1	23.23	88.16
43	$P_r + 2P_s$	$ 53P_w - P_s $	$23/43$	0.47	1.78
46	$2P_r$	$13P_w + P_s + P_r$	1	1.25	4.74
47	$ P_r - 7P_s $	$ 57P_w - P_s /17P_w + 3P_s$	$-23/47$	-0.26	-0.98

4. Power Factor Analysis of FTCP-PMVM

Under the control strategy of $I_d = 0$ and with negligible winding resistance, the stator current vector is aligned with the q-axis, while the stator voltage phasor is perpendicular to the synthetic flux linkage. Therefore, the power factor angle between the stator current and voltage is geometrically equivalent to the angle between the PM flux linkage and the total synthetic flux linkage. For this reason, the power factor can be directly derived from the geometric relationship of the flux linkage triangle. As shown in Figure 13, The PM flux linkage generated by the PMs corresponds to the active power. The reactive power originating from the inductive components consists of two parts: the air gap inductance and the leakage inductance. The apparent power is ψ , and the resulting power factor angle is φ .

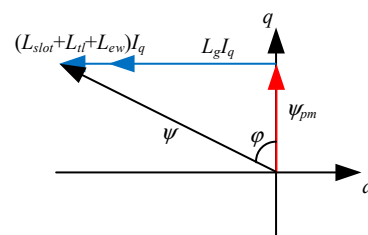


Figure 13. The phasor diagram of power factor of FTCP-PMVM.

The power factor can be expressed as:

$$PF = \cos \varphi = \frac{\psi_{pm}}{\psi} = \left[\sqrt{1 + \frac{(L_g + L_{slot} + L_{tl} + L_{ew})^2 I_q^2}{\psi_{pm}^2}} \right]^{-1} \quad (14)$$

4.1. Calculation of Inductance

4.1.1. The Model of Air Gap Inductance

The air gap self-inductance is produced by the no-load excitation current. The air gap mutual inductance is formed when the magnetic flux linkage generated by energizing

one phase couples with the winding of another phase. The self-inductance and mutual inductance can be represented by

$$\begin{cases} L_g(t) = \frac{\psi_g(t)}{I_{dc}} = \frac{\mu_0 r_g l_{st}}{g I_{dc}} \int_0^{2\pi} F_{aw, sr}(\theta, t) \cdot N_w(\theta) d\theta \\ L_m(t) = \frac{\psi_m(t)}{I_{dc}} = \frac{\mu_0 r_g l_{st}}{g I_{dc}} \int_0^{2\pi} F_{aw, sr}(\theta t) \cdot N_w(\theta \pm 2\pi/5) d\theta \end{cases} \quad (15)$$

Figure 14 illustrates the relationship between the A-phase MMF and the winding function. It can be observed that there is no interactive relationship between the different phase windings.

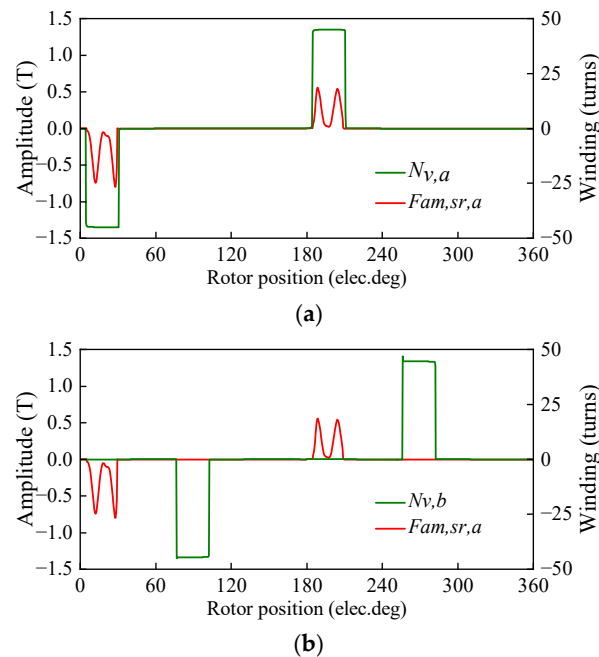


Figure 14. (a) The self-inductance and (b) mutual inductance in air gap.

4.1.2. The Model of Slot Leakage Inductance

The slots are simplified to the model shown in Figure 15. It is assumed that the leakage flux lines are straight and parallel to the slot bottom, and that the coils are uniformly distributed within the slots.

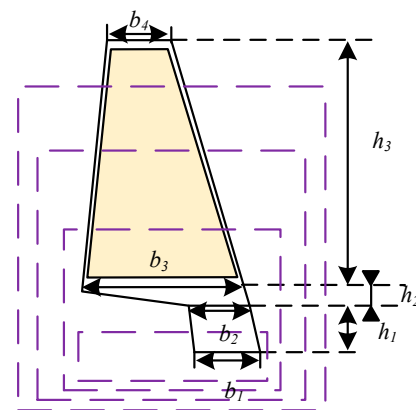


Figure 15. The simplified model of slot leakage inductance.

As shown in Figure 15, the slot leakage flux can be divided into three parts. The first part is the leakage flux in the slot opening region, corresponding to height h_1 . The second part is the leakage flux associated with the slot wedge region of height h_2 . The third part is the leakage flux linked to the winding coils, corresponding to height h_3 . The three parts of leakage flux can be expressed as

$$\begin{cases} \psi_1 = N_c^2 I_a \mu_0 l_{st} \int_0^{h_1} \frac{dh}{b_2 + (b_1 - b_2)h/h_1} \\ \psi_2 = N_c^2 I_a \mu_0 l_{st} \int_0^{h_2} \frac{dh}{b_2 + (b_3 - b_2)h/h_2} \\ \psi_3 = \left\{ \frac{[2b_4 + (b_3 - b_4)h/h_3]h}{(b_3 + b_4)h_3} \right\}^2 N_c^2 I_a \mu_0 l_{st} \int_0^{h_3} \frac{dh}{b_3 + (b_3 - b_4)h/h_3} \end{cases} \quad (16)$$

where b_1, b_2, b_3, b_4 denote the widths at different positions of stator slots. N_c represents the number of turns per phase; I_a is the amplitude of the current.

The slot leakage permeance coefficient can be expressed as

$$\begin{cases} \lambda_{s1} = \frac{\psi_{s1}}{N_c^2 I_a \mu_0 l_{st}} = \frac{h_1}{b_1 - b_2} \ln \frac{b_1}{b_2} \\ \lambda_{s2} = \frac{\psi_{s2}}{N_c^2 I_a \mu_0 l_{st}} = \frac{h_2}{b_3 - b_2} \ln \frac{b_3}{b_2} \\ \lambda_{s3} = \frac{\psi_{s3}}{N_c^2 I_a \mu_0 l_{st}} = \frac{h_3}{(b_3 + b_4)^2} \left[\frac{3b_4 + b_3}{4} + \frac{b_4^2(b_3 - 3b_4)}{2(b_3 - b_4)^2} + \frac{b_4^4 \ln(b_3/b_4)}{(b_3 - b_4)^3} \right] \end{cases} \quad (17)$$

Therefore, the total slot leakage permeance coefficient is the sum of the three components, and the slot leakage inductance can be expressed as

$$\begin{cases} \lambda_{slot} = \lambda_{s1} + \lambda_{s2} + \lambda_{s3} \\ L_{slot} = \frac{N_s}{m} N_c^2 \mu_0 l_{st} \lambda_{slot} \end{cases} \quad (18)$$

where N_s and m represent the slot number and the phase number, respectively.

4.1.3. The Model of Tooth Tip Leakage Inductance

The tooth tip leakage inductance is simplified as a regular semicircle. The leakage flux emanates from the tooth tip, traverses the air gap, reaches the rotor yoke, then crosses the air gap again, and finally arrives at the tip of the adjacent fault-tolerant tooth, as shown in Figure 16.

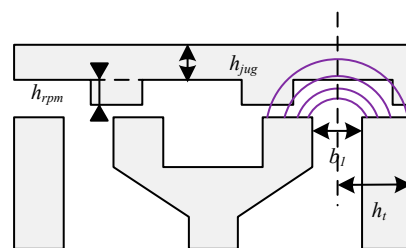


Figure 16. The simplified model of tooth tip leakage inductance.

Tooth tip leakage permeance coefficient can be expressed as

$$\lambda_{tl} = \int_{b_1/2}^{h_t} \frac{dx}{\pi x} = \frac{1}{\pi} \ln \frac{2h_t}{b_1} \quad (19)$$

Similarly, the tooth tip leakage inductance can be expressed as

$$L_{tl} = \frac{N_s}{m} N_c^2 \mu_0 l_{st} \lambda_{tl} \quad (20)$$

4.1.4. The Model of End Leakage Inductance

The end leakage inductance is the leakage inductance associated with the portions of the machine coils that extend beyond the iron core, as illustrated in Figure 17.

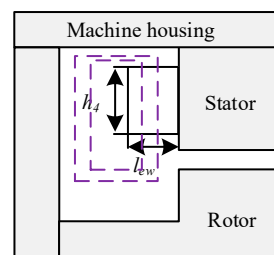


Figure 17. The simplified model of end leakage inductance.

The end winding leakage inductance can be expressed as

$$L_{ew} = \frac{N_s}{2m} N_c^2 \mu_0 \frac{\pi l_{ew}^2}{h_d} \quad (21)$$

where l_{ew} is the radius of end winding.

Therefore, the total self-inductance can be expressed as the sum of components, which can be expressed as

$$L_s = L_g + L_{slot} + L_{tl} + L_{ew} \quad (22)$$

Figure 18 shows the waveforms of self-inductance and mutual inductance. Table 6 lists the calculated values of air gap inductance, slot leakage inductance, tooth tip leakage inductance and end leakage inductance, as well as their respective contributions.

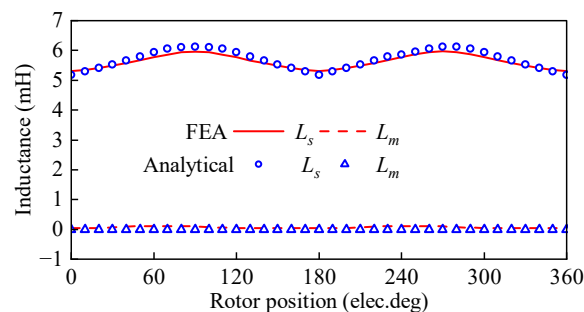


Figure 18. The waveform of total self-inductance and mutual inductance of FTCP-PMVM.

Table 6. The composition and proportion of inductance components.

Items	Value (mH)	Proportion (%)
Air gap inductance L_g	4.3	74.4
Slot leakage inductance L_{slot}	1.2	20.7
Tooth tip leakage inductance L_{tl}	0.2	3.4
End winding leakage inductance L_{ew}	0.08	1.4
L_s Includes L_{ew}	5.78	100
L_s Excludes L_{ew}	5.7	98.6
2D L_s (FEA)	5.63	97.4

4.2. Calculation of Power Factor

Figure 19 illustrates the power factor calculated by both the finite element analysis and the analytical method. It can be observed that the power factor decreases as the armature current increases. Meanwhile, the FEA results show good agreement with the analytical predictions.

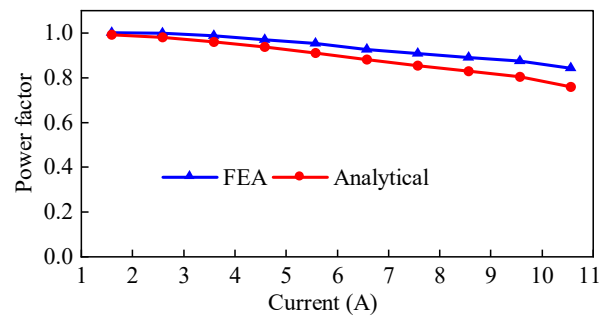


Figure 19. The relationship between power factor and current variation.

4.3. Analysis of Fault-Tolerant Performance

Figure 20 shows the torque waveforms of the FTCP-PMVM under healthy, fault-tolerant and faulty conditions. Table 7 compares the torque performance of the FTCP-PMVM under three conditions. It can be observed that the torque ripple under the fault-tolerant condition is 33.81%, which is higher than that under the healthy condition, while the average torque is only slightly lower. By contrast, the torque ripple reaches 57.12% under the faulty condition. Therefore, the machine operating in the fault-tolerant state achieves a higher average torque and much better torque ripple suppression performance compared with the faulty state. The results intuitively demonstrate the excellent fault-tolerant performance of the machine.

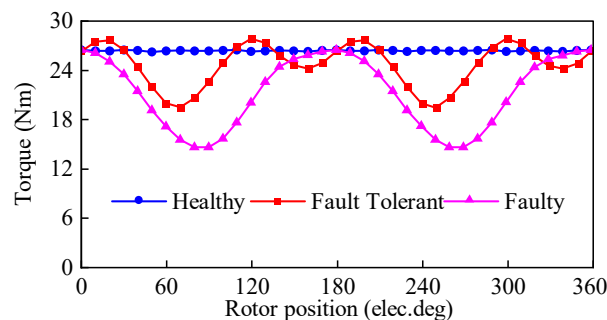


Figure 20. The torque waveforms of the FTCP-PMVM under healthy, fault-tolerant and faulty conditions.

Table 7. Comparison the machine performance under healthy, fault and fault-tolerant conditions.

Items	Healthy	Faulty	Fault-Tolerant
Average torque (Nm)	26.36	21.15	24.58
Torque ripple (%)	0.87	57.12	33.81

4.4. Comparison of SM-PMVM and FTCP-PMVM

Figure 21 illustrates the topologies of the surface-mounted permanent-magnet Vernier machine (SM-PMVM) with 20 slots and 31 pole pairs, and the FTCP-PMVM with 20 slots and 23 pole pairs. Detailed design parameters of the two machines, such as the magnet pole arc coefficient and magnet thickness, are listed in Table 8 which compares their electromagnetic performance. It can be observed that compared with the conventional SM-PMVM, the proposed FTCP-PMVM consumes only 53.36 cm³ of the permanent-magnet material, while achieving an average torque of 26.36 Nm. Therefore, the FTCP-PMVM topology not only reduces the usage of permanent magnets significantly but also improves the average output torque and suppresses torque ripple.

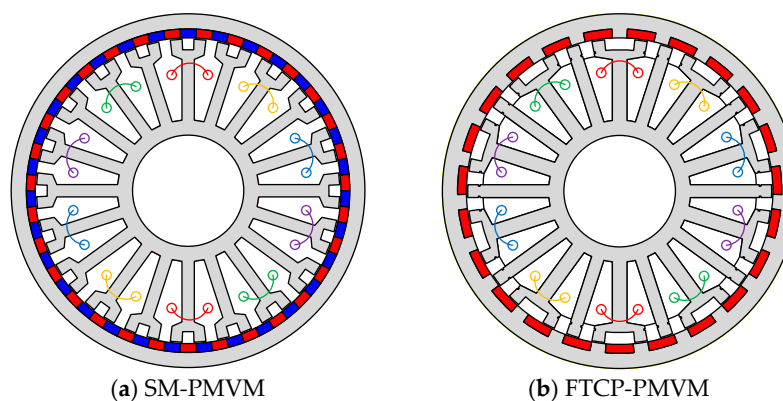


Figure 21. Topological structure of (a) SM-PMVM and (b) FTCP-PMVM.

Table 8. Comparison performance of SM-PMVM and FTCP-PMVM.

Items	SM-PMVM	FTCP-PMVM
PM pole pairs	31	23
Pole arc coefficient of PM	1	1.3
Stator slot number		20
PM thickness (mm)		3.5
Remanence flux density of Permanent magnet (T)		1.23
Torque density (Nm/L)	25.24	28.54
PM volume (cm ³)	82.04	53.36
Power factor	0.56	0.86
Torque ripple (%)	5.12	0.87
Average torque (Nm)	23.32	26.36

5. Experimental Verification

To verify the accuracy of the aforementioned analytical models for electromagnetic torque and power factor, a prototype model of the FTCP-PMVM is built in this paper. The rotor lamination contains 23 slots for permanent magnets, while the stator is composed of alternating fault-tolerant teeth and split teeth, as shown in Figure 22. The corresponding experimental environment is illustrated in Figure 23, which primarily consists of five parts: the platform, the control system, the record system, the power supply system and the sensor modules. The experimental platform consists of a prototype machine, torque sensor, magnetic powder brake, tension controller and optical encoder. The torque sensor measures the output torque. The magnetic powder brake and tension controller jointly apply mechanical load to the machine. The optical encoder collects real-time rotor position and speed signals to provide position feedback. The control board drives the prototype with five-phase sinusoidal current output from the inverter. The power supply system supplies power to the inverter and torque sensor. An oscilloscope records the experimental waveforms.

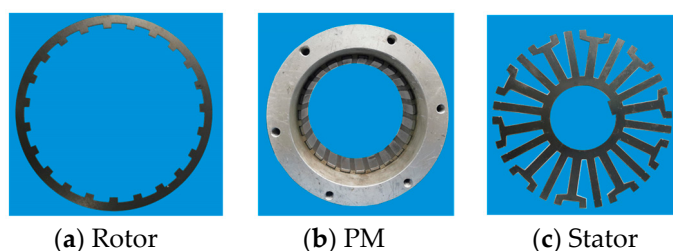


Figure 22. Prototype machine.

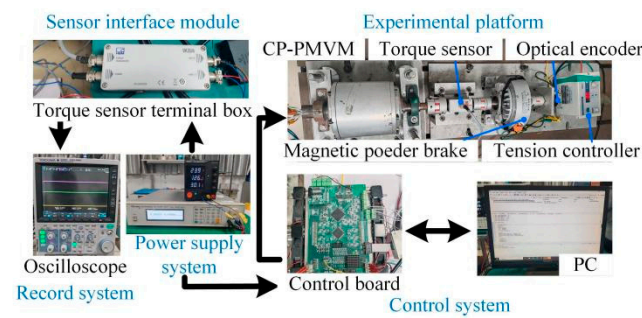


Figure 23. Experimental system.

Figure 24 provides the no-load back-EMF waveforms and harmonic spectra by finite element analysis and experiment at the speed of 600 r/min. It can be found that the experimental back-EMF waveform is close to that of finite element analysis. The error of the fundamental components of the no-load back-EMFs between them is small, which mainly results from the mechanical error and neglected end effect in two-dimensional finite element analysis.

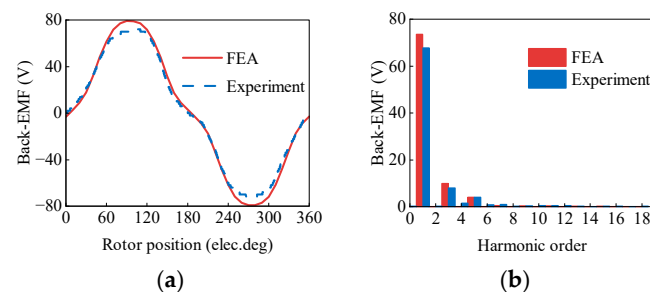


Figure 24. No-load back-EMF by experiment and finite element analysis. (a) Waveforms. (b) Harmonic spectra.

Figure 25 depicts a comparison of the torque performance of the FTCP-PMVM under different input currents, including measured, analytical and finite element analysis values. It can be observed that the electromagnetic torque increases as the current rises. Furthermore, the finite element analysis results are slightly higher than the experiment ones, which is consistent with the back-EMF analysis shown in Figure 24. Figure 26 compares the power factors obtained from experiments and simulations under different current levels. The measured results show good agreement with the simulated data, with an error of approximately 10% between the two sets of values. This discrepancy is mainly caused by stray losses, manufacturing tolerances, end leakage inductance and additional impedance of the experimental platform.

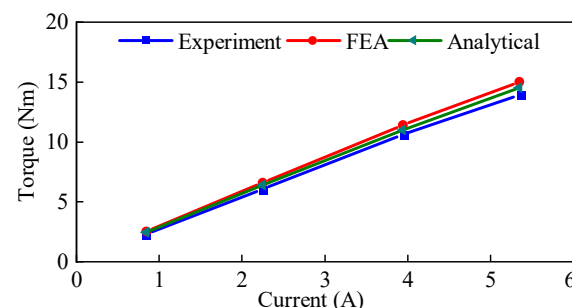


Figure 25. Comparison of electromagnetic torque under different input currents.

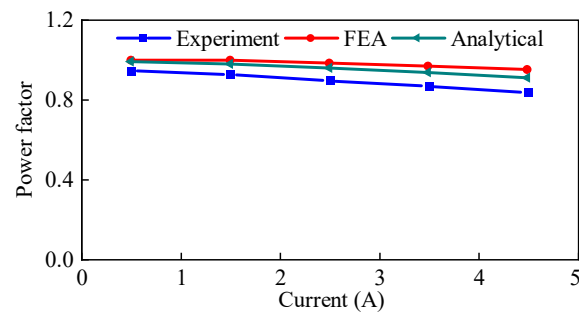


Figure 26. Comparison of power factor under different input currents.

To further evaluate the dynamic performance of the consequent-pole permanent-magnet Vernier machine, variable-load experiments were carried out, as shown in Figure 27. When the load torque was stepwise varied from 0 to 14 Nm, no obvious drop or overshoot was observed in the regulation of the output torque. Furthermore, during the stepwise increase in the load, the machine speed remained stable at 400 r/min, while the phase current increased synchronously with the load, with the root mean square current value varying from 0 to 4 A. The current waveforms were smooth and exhibited no significant distortion. Figure 28 shows the efficiency map of the FTC-PPMVM, where all efficiency values are calculated based on measurements from the power analyzer. It can be observed that within the tested operating range, the efficiency increases gradually with the rising speed and torque.

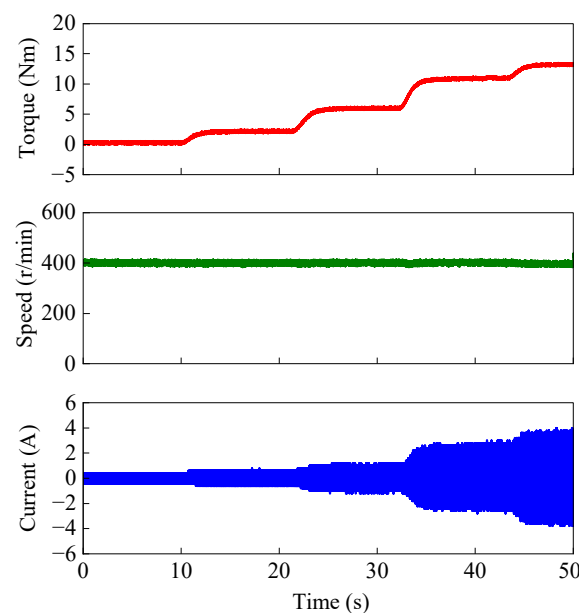


Figure 27. The waveforms of machine variable-load experiment.

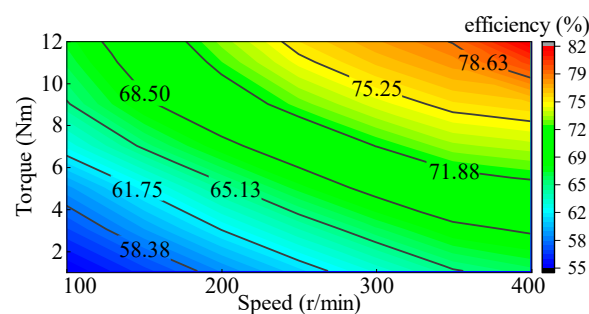


Figure 28. Efficiency map of FTCP-PMVM.

6. Conclusions

In this study, a comprehensive analysis of the torque generation mechanism and power factor for the FTCP-PMVM is presented, grounded in the theory of magnetic field modulation. Given the double-salient structure of the FTCP-PMVM, the abundant harmonics in both the PM and armature fields are investigated. By matching harmonics with the same order and rotational speed, the individual torque contributions of various harmonic components are effectively quantified. Furthermore, the key inductance parameters, including self, mutual and leakage inductances, are analytically derived to clarify their impact on the power factor. The results indicate that the FTCP-PMVM possesses high self-inductance and low short-circuit current, combined with low mutual inductance and minimal inter-phase interference. Consequently, the FTCP-PMVM exhibits excellent fault-tolerant capability. The power factor obtained from FEA shows good agreement with the results from the analytical model, thereby demonstrating the effectiveness and accuracy of the proposed approach. Finally, an experimental prototype and a test platform are established. Experimental results exhibit great consistency with theoretical predictions, which further verifies the reliability of the proposed analytical framework.

Author Contributions: Project administration, T.J.; experiment, P.D.; visualization, C.G.; writing—original draft, W.B. All authors have read and agreed to the published version of the manuscript.

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Conflicts of Interest: The authors declare no conflicts of interest.

Abbreviations/Symbols

The following are some of the symbols and abbreviations used in the text:

Symbols	
P_r	PM pole pair
P_s	number of stator tooth unit
θ	represent mechanical angle
Ω_r	mechanical angular velocity
$a_{r,0}, a_{s,0}$	coefficients of direct current component
$a_{r,j}, a_{s,j}$	coefficients of harmonic component
$F_{pm,k}$	Fourier coefficient of PM MMF harmonic
μ_0	vacuum permeability
g	length of air gap
P_w	Pole pair number of windings
I_{rms}	root mean square value of the current
r_g	air gap radius
l_{st}	stack length
B_r	radial air gap magnetic flux density
B_t	tangential air gap magnetic flux density
N_s	slot number
N_c	number of turns per phase
m	phase number
L_s	self-inductance

L_g	air gap self-inductance
L_{slot}	slot leakage inductance
L_{tl}	tooth tip leakage inductance
L_{ew}	end winding leakage inductance
L_m	mutual inductance
Abbreviations	
MMF	Magnetomotive Force
EMF	Electromotive Force
SPM	Surface Permanent Magnet
FEA	Finite Element Analysis

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