

OPTIMIZATION OF THE ELECTROMAGNETIC PERFORMANCE OF AXIAL-FLUX PERMANENT MAGNET MOTORS FOR MEDIUM- AND LARGE-SIZED UAVs

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Abstract

To address the demand for high power density and lightweight designs in medium-to-large-sized unmanned aerial vehicles (UAVs), this paper presents a dual-rotor, single-stator axial flux permanent magnet motor (AFPM). Based on three-dimensional finite element analysis, the severe eddy current losses caused by spatial higher harmonics were effectively suppressed through the coordinated optimization of the radial three-block configuration of the permanent magnets and the pole arc coefficient. Following optimization, eddy current losses in the permanent magnets were reduced from 222 W to 70 W, the overall motor efficiency was increased to 92%, the average output torque was increased to 59.1 Nm, and torque ripple was reduced to 4.5%.

1 Introduction

With the advancement of aviation electrification technology, the demand for high-power-density motors in medium- and large-sized UAVs is becoming increasingly urgent. Reference [1] compared the performance of axial-flux and radial-flux motors, showing that traditional radial-flux motors are often limited by radial dimensions and thermal dissipation bottlenecks when increasing power density. In contrast, the axial-flux permanent magnet motor (AFPM), with its compact disc-shaped structure, extremely short axial dimensions, and high torque density, has emerged as an ideal choice for direct-drive propulsion systems in UAVs [2]. Based on this conclusion, Reference [3] designed a novel single-stator, single-rotor AFPM and demonstrated its suitability for UAV applications through three-dimensional finite element analysis. References [4] and [5] conducted comparative analyses of various topological structures. The results indicate that the yoke-less stator AFPM (YASA-AFPM) maintains high power density while possessing the ability to withstand significant axial forces, making it better suited for the diverse operational requirements of medium-to-large UAVs. At the theoretical design level, References [6] and [7] systematically analyzed methods for determining key parameters, providing a theoretical foundation for the design of the main body; Reference [8] focuses on the topological optimization of rotor permanent magnets, broadening the scope for improving electromagnetic designs. Based on the above research, this paper designs an AFPM suitable for medium-to-large UAVs. After completing the derivation of basic electromagnetic dimensions and structural design, the study focuses on investigating the generation mechanism of higher-order harmonics in the air gap and their impact on

eddy current losses in permanent magnets. Through three-dimensional finite element analysis, the practical effectiveness of radial segmenting of permanent magnets and optimization of pole arc parameters in suppressing losses is evaluated, achieving an optimal balance between high torque output and low eddy current losses in the motor.

2 Structural Parameters and Electromagnetic Principles

To meet the high power density requirements of medium- and large-sized drones, this design employs a YASA-AFPM topology. By eliminating the stator yoke to form a closed magnetic circuit with two rotors and a single stator, it balances the axial electromagnetic forces and reduces core losses. The structural model is shown in Figure 1.

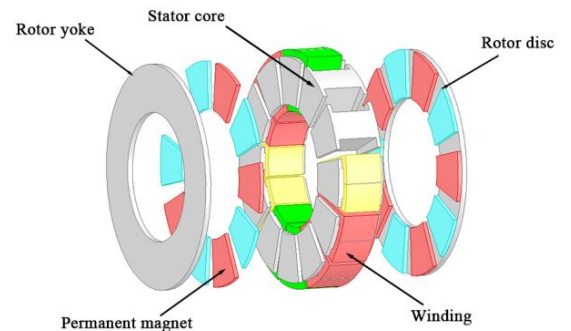


Fig. 1 YASA-AFPM Structural Model

2.1 Design of Windings, Pole Count, and Slot Count

This design employs a 10-pole, 12-slot configuration with a fractional-slot concentrated winding arrangement. The winding pitch is 1, and the fundamental winding coefficient is 0.933, resulting in high fundamental utilization. The least common multiple of the pole and slot numbers is 60, which allows for high output capacity while keeping the vibrations and noise caused by cogging torque within reasonable limits. Figure 2 shows the three-phase winding layout.

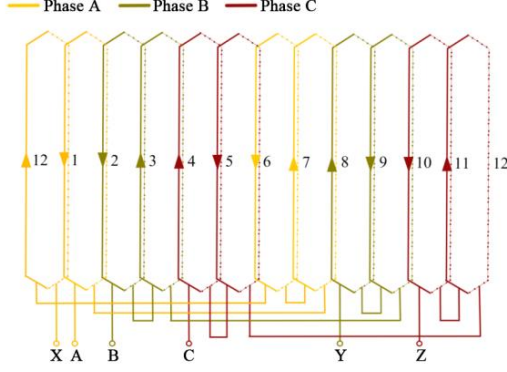


Fig. 2 The three-phase windings of an electric motor

2.2 Determination of Stator and Rotor Parameters

The YASA-AFPM magnetic circuit model is shown in Figure 3. The main magnetic flux starts at the N pole on the lower rotor disc, travels axially through the first air gap to the stator core; It then departs from the stator core, passes through the second air gap, and reaches the S pole on the upper rotor disc; subsequently, it flows through the yoke of the upper rotor disc to the adjacent N pole; then, it travels in the opposite direction through the second air gap, the stator core, and the first air gap, returning to the S pole adjacent to the lower N pole, and finally returns along the rotor yoke to the original N pole, forming a closed circuit [9].

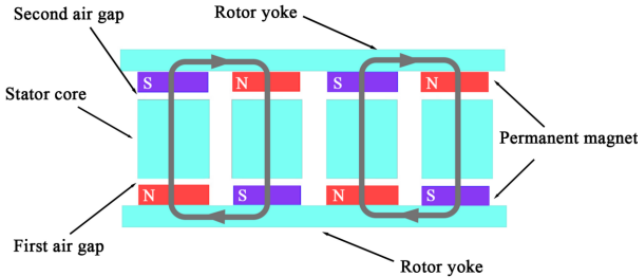


Fig. 3 Magnetic circuit model

Based on the magnetic circuit model described above, analyze the magnetic flux density in the air gap of the magnetic circuit. Derive the relationship between the magnetic field in the air gap and the total magnetic flux.

$$H_{\delta}\delta = 2H_m h_m \quad (1)$$

$$\phi_m = \phi_{\delta} + \phi_{\sigma} \quad (2)$$

where δ is the total air gap length, h_m is the length along the magnetization direction of the permanent magnet, H_{δ} is the magnetic field strength in the air gap, H_m is the magnetic field

strength of the permanent magnet, ϕ_m is the total magnetic flux of the YASA-AFPM, ϕ_{δ} is the main magnetic flux, and ϕ_{σ} is the leakage flux. By further introducing the leakage coefficient σ and applying the principle of magnetic flux continuity, we obtain

$$A_m B_m = \sigma A_{\delta} B_{\delta} \quad (3)$$

Here, A_m represents the magnetic flux area per pole under the permanent magnet, B_m represents the magnetic flux density of the permanent magnet, A_{δ} represents the effective area of the air gap, and B_{δ} represents the magnetic flux density in the air gap. To simplify the calculation, let the pole arc coefficient be equal to the calculated pole arc coefficient, and let the air gap magnetic flux density distribution coefficient be 1. The following equation for calculating the air gap magnetic flux density can then be derived

$$B_{\delta} = \frac{2B_r}{2\sigma + \mu_r} \frac{\delta}{h_m} \quad (4)$$

Here, B_r is the remanence of the permanent magnet, and μ_r is the relative magnetic permeability of the permanent magnet material. The axial electromagnetic force generated by a single air gap is

$$F_a = \frac{B_{\delta}^2 \cdot A_m}{2\mu_0} \quad (5)$$

where μ_0 is the magnetic permeability of air; substituting Equation (4) into the axial force equation yields

$$F_a(\delta) \propto \frac{1}{(h_m + \mu_r \delta)^2} \quad (6)$$

From the above formula, it can be deduced that an increase in the total air gap length reduces the magnetic flux density in the air gap, which in turn causes a decrease in torque; conversely, a decrease in the air gap length leads to an increase in axial force, affecting the motor's stability. Taking these factors into account, the first air gap length is set to 1 mm, the second air gap length to 1 mm, and the total air gap length to 2 mm. The preliminary dimensions of the YASA-AFPM are as follows

Table 1 YASA-AFPM Specifications

Motor Specifications	value	Motor Specifications	value
Rated Power/kW	14	Groove width/mm	7.19
Rated torque/Nm	56	Rotor core thickness/mm	4
Rated speed/rpm	2400	Permanent magnet thickness/mm	3
Pole-slot combination	12s10p	Number of parallel branches	2
Motor outer diameter/mm	165	Pole embrace	0.72
Motor inner diameter/mm	95	Number of conductors per slot	120

Motor Specifications	value	Motor Specifications	value
Total air gap length/mm	2	Stator core length /mm	36

3 Finite Element Simulation Analysis

3.1 No-Load analysis

The motor was modeled and analyzed using Ansys finite element simulation software. With the current excitation set to zero, an induced electromotive force was generated in the stator windings as the rotor rotated. By analyzing the waveform of the induced electromotive force, the reasonableness of the design was preliminarily assessed. Figure 4 shows the no-load induced EMF waveform. As shown in the figure, the root mean square (RMS) values for phases A, B, and C are 267.16 V, 277.42 V, and 279.64 V, respectively. The three-phase average RMS value is 274.74 V, and the three-phase asymmetry is 2.76%. Figure 5 shows the amplitudes of each order obtained by Fast Fourier Transform(FFT) of the induced EMF in Phase A. As shown in the figure, the fundamental amplitude of the Phase A induced EMF waveform is 386.76, and the third harmonic has a relatively high amplitude compared to other harmonic orders, at 13.64.

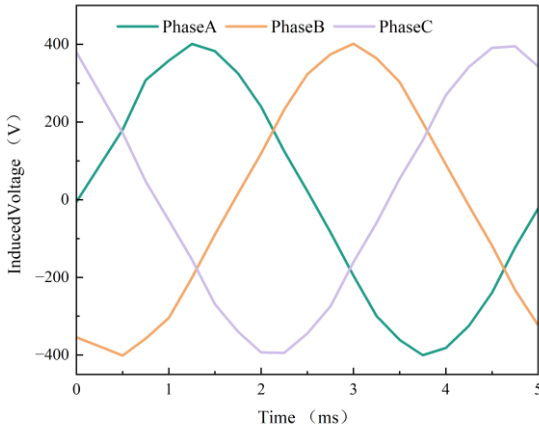


Fig. 4 Waveform of No-Load Induced EMF

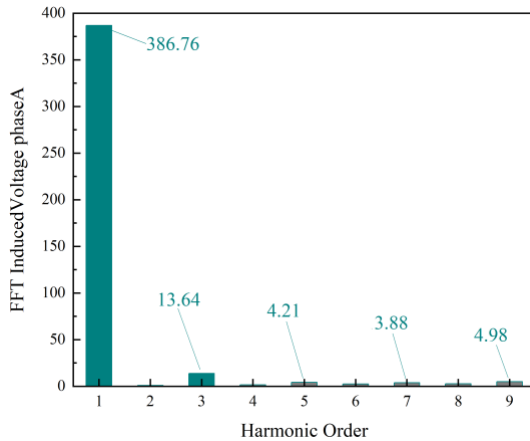


Fig. 5 Phase A induced emf FFT

Under no-load conditions, the stator slots cause non-uniform magnetic permeability in the air gap, thereby generating tooth harmonics in the air gap. These tooth harmonics induce eddy currents in the rotor's permanent magnets, resulting in eddy current losses. Figure 6 shows the air gap magnetic flux density waveform. As can be seen from the figure, the overall air gap magnetic flux density exhibits an approximate trapezoidal wave with a flat top, and there are sharp high-frequency fluctuations at the wave crest, reflecting the significant tooth slot effect caused by the stator slots. Figure 7 shows the amplitudes of each harmonic order obtained after FFT of the air gap magnetic flux density waveform. The fundamental amplitude is 0.70, and the 5th harmonic amplitude is 0.17, accounting for 24.6% of the fundamental.

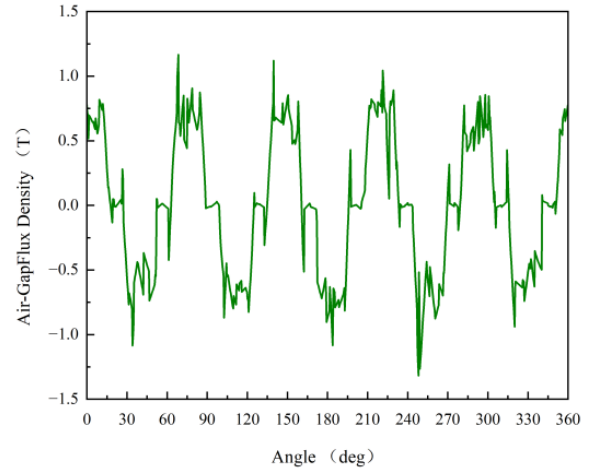


Fig. 6 Air Gap Magnetic Flux Density Waveform

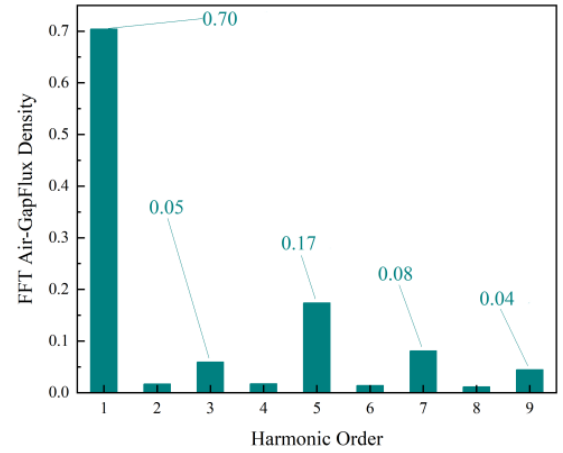


Fig. 7 Air Gap Magnetic Flux Density Waveform FFT

3.2 Load analysis

Load analysis effectively verifies whether the motor's overall performance meets engineering requirements by evaluating the degree of induced electromotive force distortion, the magnitude of output torque, and loss characteristics. By controlling the excitation current at 20 A, the output torque was determined, as shown in Figure 8. As can be seen from the figure, the average output torque is 57.94 Nm, with a torque ripple of 6.2%, which meets the performance requirements. The loss waveforms are shown in Figure 9. Analysis of the graph reveals that under rated

operating conditions, the copper losses in the windings, iron losses in the stator and rotor, and eddy current losses in the permanent magnets are 860.03 W, 56.43 W, and 222 W, respectively. Copper losses account for the largest proportion, while iron core losses are relatively low due to the moderate rated speed. Based on design experience, the motor's mechanical losses and additional losses are set at 1.00% of the output power. The motor efficiency calculated for this initial design is 90.8%.

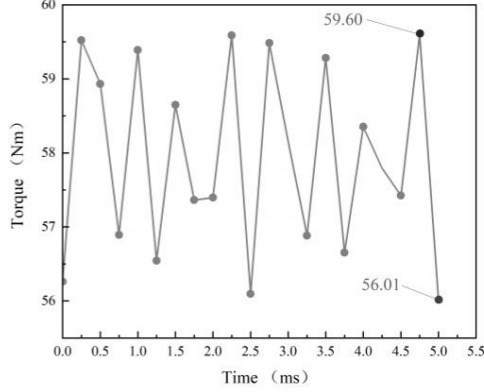


Fig. 8 Output torque

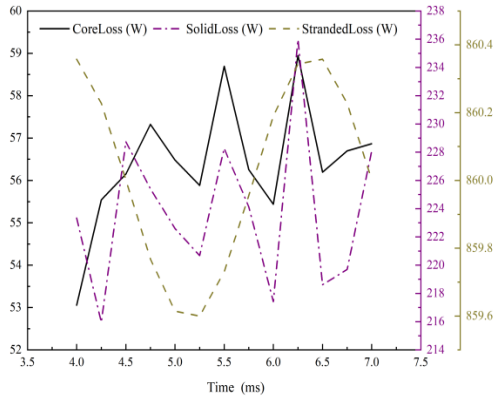


Fig. 9 Load loss waveform

3.3 Structural Optimization Topology

By radially segmenting the permanent magnets on the rotor, the circulation path of eddy currents is interrupted, increasing the equivalent resistance of the eddy current loop. This breaks down a large eddy current loop into multiple small ones, resulting in a significant reduction in eddy current losses. The segmented model is shown in Figure 10.

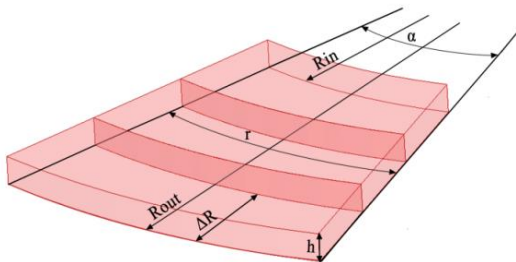


Fig. 10 Radial Partitioning Model

Let the circumference of the circle be r , the inner diameter of the permanent magnet sector be R_{in} , the outer diameter be R_{out} , the pole arc angle be α , the material resistivity be ρ , and the axial magnetic flux density be B_z ; the circumferential electric field strength E_θ satisfies [11]

$$\frac{1}{r} \frac{\partial(rE_\theta)}{\partial r} = -\frac{\partial B_z}{\partial t} \quad (7)$$

Integrating over r yields the circumferential electric field intensity at any radius r

$$E_\theta(r) = -\frac{1}{2} r \frac{\partial B_z}{\partial t} \quad (8)$$

Consider a circumferential element with a radial thickness of dr and an axial thickness of h at radius r . Neglecting the axial length of the circumferential loop, let the path length of the circumferential loop be $2\alpha r$. Introduce an end-fold correction factor β_0 , and integrate over the circumferential loop to obtain the induced emf element $\mathcal{E}(r)$ and the equivalent resistance $R(r)$ of the loop.

$$\mathcal{E}(r) = -\alpha r^2 \frac{\partial B_z}{\partial t} \quad (9)$$

$$R(r) = \beta_0 \cdot \rho \frac{2\alpha r}{h \cdot dr} \quad (10)$$

The eddy current losses power element dP is

$$dP = \frac{\mathcal{E}(r)^2}{R} \quad (11)$$

By combining equations (9), (10), and (11) and integrating over the entire interval $[R_{in}, R_{out}]$, we obtain the eddy current losses P_{total} of the permanent magnet when it is not segmented.

$$P_{total} = \frac{h\alpha}{8\beta_0\rho} \left(\frac{\partial B_z}{\partial t} \right)^2 (R_{out}^4 - R_{in}^4) \quad (12)$$

If the permanent magnet is divided radially into N equal segments, the radial width ΔR of each segment is

$$\Delta R = \frac{R_{out} - R_{in}}{3} \quad (13)$$

By introducing the narrowband two-dimensional correction factor β_{seg} and summing the losses of each independent loop, the eddy current losses power P_{seg} after segmentation is obtained as

$$P_{seg} \approx \frac{h\alpha}{72\rho \cdot \beta_{seg}} \left(\frac{\partial B_z}{\partial t} \right)^2 (R_{out} - R_{in})^3 (R_{out} + R_{in}) \quad (14)$$

By combining equations (12) and (14), we obtain the following equation for comparing total eddy current losses

$$P_{\text{seg}} \approx \frac{1}{N^2} \cdot \left(\frac{\beta_0}{\beta_{\text{seg}}} \right) \cdot \left[\frac{(R_{\text{out}} - R_{\text{in}})^2}{R_{\text{out}}^2 + R_{\text{in}}^2} \right] \cdot P_{\text{total}} \quad (15)$$

As shown by the equation, the greater the number of radial segments, the lower the eddy current losses in the permanent magnet. Therefore, this design divides the permanent magnet into three equal segments. Figure 11 shows a comparison of the eddy current losses in the YASA-AFPM before and after segmentation. The average eddy current losses before optimization was approximately 222 W. By segmenting the permanent magnet, the average eddy current losses was reduced to approximately 70 W, a decrease of about 68% compared to the pre-optimization value, resulting in an overall efficiency of 92%. Figure 12 shows a comparison of the output torque before and after the permanent magnet optimization. As shown in the figure, the optimized output torque ranges from 57.76 Nm to 60.44 Nm, with an average torque of 59.16 Nm and a torque ripple of 4.5%.

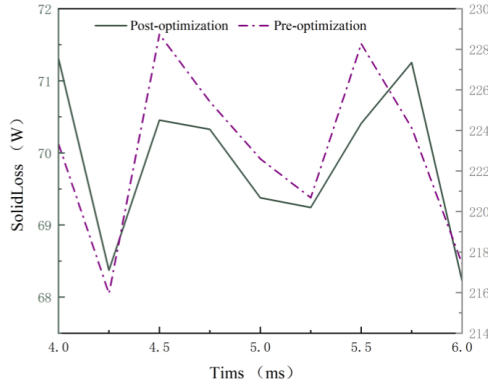


Fig. 11 Comparison of Eddy Current Losses

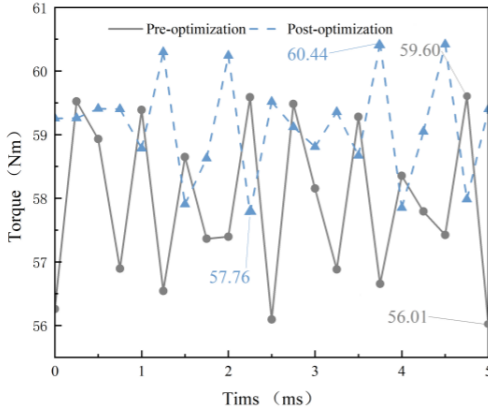


Fig. 12 Comparison of Output Torque

Adjusting the polar arc coefficient can effectively improve the air gap magnetic flux density waveform. As shown in Reference [12], the Fourier expansion coefficient M_v of the magnetization intensity M of a radially magnetized permanent magnet is

$$M_v = 2(B_r / \mu_o) \alpha_p \frac{\sin\left(\frac{v\pi\alpha_p}{2}\right)}{\frac{v\pi\alpha_p}{2}} \quad (16)$$

Here, v represents the spatial harmonic order, α_p is the pole arc coefficient, B_r is the residual magnetic flux density, and μ_o is the vacuum magnetic permeability. Since the magnetization M acts as a source field to induce the air-gap magnetic flux density B_δ in the permanent magnet, the amplitudes B_v of the spatial harmonics of the air-gap magnetic flux density also follow the same proportional relationship as M_v . Therefore, the relationship between the amplitude of the spatial harmonic order in the air-gap magnetic flux density and the pole arc coefficient can be expressed as

$$B_v \propto \frac{1}{v} \sin\left(v \cdot \alpha_p \cdot \frac{\pi}{2}\right) \quad (17)$$

As can be seen from the equation, the amplitude of the spatial harmonics in the air gap magnetic flux density is approximately proportional to the pole arc coefficient. Therefore, by performing a finite element parameter scan, the circumferential width of the permanent magnet can be adjusted to reduce the 5th harmonic. Figure 13 shows the results of the FFT order analysis of the air gap magnetic flux density for different pole arc coefficients. As shown in the figure, as the pole arc coefficient increases from 0.64 to 0.78, the amplitude of the fundamental component exhibits an overall upward trend, reaching a peak of approximately 0.86 when the pole arc coefficient is 0.76. Meanwhile, odd harmonics such as the 3rd, 5th, and 7th harmonics are attenuated at pole arc coefficients of 0.68, 0.70, and 0.74, respectively. Figure 14 shows the dynamic waveform of the electromagnetic torque versus time for the YASA-AFPM under different pole-to-pole ratios. As shown in the figure, the motor's output torque increases with the pole-to-pole ratio, reaching a peak of approximately 59.37 Nm at a pole-to-pole ratio of 0.70. By comprehensively comparing the performance data from Figures 13 and 14, as shown in the comparison table in Table 2, and based on the information in the table, combined with the analysis of the air gap magnetic flux density discussed above, and considering the high 5th harmonic at a pole arc coefficient of 0.70, the optimized pole arc coefficient was ultimately selected as 0.68.

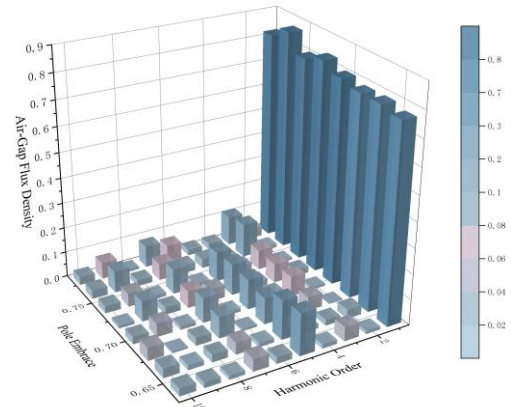


Fig. 13 Comparative FFT Analysis of Air-Gap Magnetic Flux Density

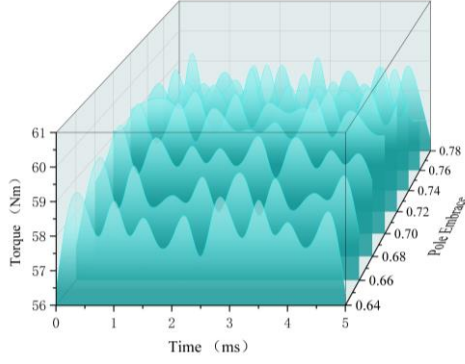


Fig. 14 Comparative FFT Analysis of output torque

Table 2 Comparative Analysis of Harmonic Orders

Pole Embrace	Harmonic Orders				Output Torque / Nm
	1	3	5	7	
0.68	0.78	0.02	0.10	0.09	59.06
0.7	0.77	0.06	0.12	0.08	59.24
0.72	0.78	0.07	0.08	0.09	59.84
0.74	0.84	0.14	0.05	0.11	58.77
0.76	0.83	0.11	0.07	0.06	58.36

4 Conclusion

This paper presents the design and optimization of a 10-pole, 12-slot YASA-AFPM motor tailored to the requirements of medium-to-large unmanned aerial vehicles (UAVs). The design employs a split-slot concentrated winding configuration with a fundamental winding coefficient as high as 0.933. It ensures an initial high torque output of 57.9 Nm while effectively suppressing torque ripple and cogging torque. Three-dimensional finite element simulations were conducted to analyze the motor's no-load and load transient field behavior. The results indicate that stator cogging effects cause eddy current losses in the rotor-side permanent magnets as high as 222 W. To address this issue, this paper effectively breaks the macroscopic eddy current loop by radially segmenting the permanent magnets, reducing eddy current losses to 70 W. Combined with parameter scanning, the pole arc coefficient was optimized to 0.68, effectively suppressing the 5th harmonic while increasing the average output torque to 59.1 Nm. The optimized motor achieves a total efficiency of 92%, meeting the comprehensive requirements of medium- and large-sized UAVs for high efficiency, high torque, and high reliability in electric drive systems. This provides an important theoretical basis and engineering reference for the low-loss design of high-power-density axial flux motors for medium- and large-sized UAVs.

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