

Article

Constraint-Based Multi-Pole EL Map Screening for Pole and Slot Selection in Axial Flux Motors of Collaborative Robot Joints

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Abstract

This paper proposes an inverter constraint-based EL map screening method for selecting pole–slot combinations of an axial flux permanent magnet motor (AFPM) for collaborative robot joints according to the requirements of the actual drive system. First, an initial set of candidates is identified based on the winding factor and pole and slot characteristics. The inductance and no-load back-EMF of each candidate are then represented on a common EL map, enabling different pole and slot topologies to be compared within the same electromagnetic parameter space. Subsequently, the required current and voltage for generating the target torque are calculated and evaluated against the inverter voltage and current limits to assess the system-level drive suitability of each candidate. Among the candidates satisfying both the voltage and current constraints, the current margin is used as the primary selection criterion, while the harmonic characteristics of the no-load back-EMF are additionally considered. Based on this evaluation, the 22-pole and 24-slot combination is selected for detailed design. For the selected AFPM, a detailed 3D finite element method (FEM) design is performed considering the number of turns, permanent magnet length, magnet spacing, and tooth spacing as design variables. Under the same motor volume constraint as the conventional radial flux permanent magnet motor (RFPM), the final AFPM increases the load torque from 0.737 N·m to 0.907 N·m and the output power from 263.6 W to 332.4 W, corresponding to improvements of 23.1% in torque density and 26.1% in power density, respectively.

Keywords: axial flux permanent magnet motor; collaborative robot; pole–slot combination; inverter constraint; power density

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1. Introduction

Recent years have witnessed a rapid expansion in the application of collaborative robots (cobots) in manufacturing, accompanied by substantial growth in the related market [1,2]. Collaborative robots are designed to operate in the same workspace as human workers and can therefore be deployed without separate safety fences. This can improve workplace safety while reducing facility and operating costs and alleviating labor shortages [1,2]. Under these operating conditions, compact and lightweight robot bodies and drive systems are important for ensuring both worker safety and robot maneuverability.

At the same time, joint actuators must provide sufficiently high torque within a limited volume to achieve a high payload capacity. In addition, internal wiring structures, in which power and signal cables are routed through the center of the joint, are commonly adopted to reduce interference and safety concerns associated with external cables. Consequently, drive motors with a large diameter hollow shaft structure are required.

Under these constraints, an axial flux permanent magnet motor (AFPM) is a promising alternative to a conventional radial flux permanent magnet motor (RFPM). Owing to its disk-shaped structure with a short axial length relative to its diameter, an AFPM can provide a relatively large internal hollow region, which is advantageous for routing cables through robotic joints [3,4]. In addition, the large effective air gap area and torque-producing radius associated with its larger diameter can provide structural advantages for improving torque density under the same motor volume constraint [5,6]. Therefore, in this study, an existing RFPM used in a collaborative robot joint is adopted as the baseline model, and an AFPM is designed to replace the RFPM while maintaining the same overall motor volume and improving the power density.

Because the electromagnetic geometry of an AFPM varies along the radial direction and the magnetic flux distribution is inherently three-dimensional, three-dimensional finite element method (3D FEM) analysis is widely used for accurate electromagnetic performance evaluation [7,8]. To reduce the associated analytical burden and improve design efficiency, various AFPM design approaches based on analytical models, quasi-3D models, magnetic equivalent circuits (MECs), and optimization techniques have been investigated [8,9]. Hemeida et al. proposed a quasi-3D MEC model that accounts for the three-dimensional effects of a surface-mounted AFPM through radial multi-slicing, enabling efficient prediction of electromagnetic characteristics such as no-load and load voltages, torque ripple, and core loss [9]. Benlamine et al. presented a design approach that optimizes the main design variables of an AFPM using a simplified analytical model combined with a genetic algorithm [10]. These studies provide useful approaches for efficiently predicting AFPM electromagnetic characteristics and optimizing detailed geometric parameters.

The performance of an AFPM is also strongly affected by its pole–slot combination. Variations in the numbers of poles and slots influence the winding factor, back-EMF, inductance, magnetomotive force harmonics, and electromagnetic force distribution. Previous studies have investigated pole–slot combinations and concentrated winding configurations from the perspectives of inductance characteristics, winding factor, electromagnetic performance, and winding periodicity [11–14]. Therefore, selecting an appropriate pole and slot topology is an important step in the preliminary design of an AFPM.

However, when pole–slot combinations are compared solely using motor-side indices such as the winding factor, greatest common divisor (GCD), back-EMF, or individual electromagnetic characteristics, it is difficult to directly determine whether a given topology can generate the required target torque within the voltage and current limits of the actual drive inverter. In particular, a high back-EMF is advantageous in reducing the current required to generate a given torque, but it may simultaneously increase the required voltage. In addition, the reactance voltage determined by the inductance and electrical angular speed also affects the inverter voltage requirement. Therefore, pole and slot selection requires a system-level selection procedure that considers not only individual motor characteristics but also the target torque and the voltage and current constraints of the drive inverter.

To address this issue, this study proposes an inverter constraint-based EL map screening method for preliminary pole and slot topology selection. The no-load back-EMF

and inductance obtained from 3D FEM are represented in a common electromagnetic parameter space and linked to the target torque and inverter voltage/current constraints. Based on this screening, a suitable pole and slot topology is selected and subsequently subjected to detailed 3D FEM design.

2. Design Methodology

Figure 1 illustrates the overall design procedure of the AFPM proposed in this paper. First, the required specifications and design constraints of the motor for collaborative robot joint actuation are defined, and the existing RFPM is selected as the baseline model. Then, to achieve high torque and output within a limited volume, the winding factor and pole and slot characteristics are preliminarily evaluated based on theoretical equations, and an initial set of pole–slot candidates is derived.

For each candidate, the geometry is established under identical initial design conditions, and the no-load back-EMF and inductance are obtained using 3D FEM. The back-EMF and inductance of each candidate are then represented on a common EL map, and the required current and voltage for generating the target torque are calculated. By comparing the calculated required current and voltage with the inverter current and voltage constraints, the system-level drive suitability of each pole–slot combination is evaluated. Among the candidates satisfying the voltage and current constraints, the final pole and slot topology for detailed design is selected by additionally comparing the current margin remaining after generating the target torque and the harmonic characteristics of the no-load back-EMF.

For the topology selected using the stepwise EL map-based evaluation, a detailed 3D FEM design is subsequently performed. In this process, key design variables such as the number of turns, permanent magnet length, permanent magnet spacing, and tooth spacing are varied, and their effects on the electromagnetic performance and magnetic saturation characteristics are analyzed. Finally, the AFPM geometry is determined to achieve high output performance while satisfying both the inverter voltage/current constraints and the magnetic saturation constraint.

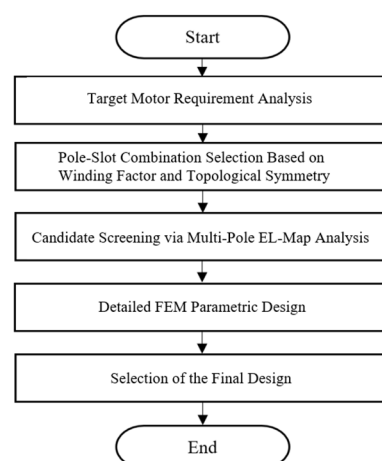


Figure 1. Design procedure for the axial flux permanent magnet (AFPM) motor.

2.1. Target Motor Analysis

The 20-pole/18-slot RFPM considered in this study is the existing motor used in the target collaborative robot joint and was not newly selected or optimized for the present comparison. Therefore, the RFPM is used as an application-specific baseline that defines

the required operating point and dimensional envelope. The baseline RFPM employs a SPM configuration. A common air gap length of 0.6 mm was imposed as a reference condition for the present replacement design. Accordingly, the comparison in this study is intended to evaluate the proposed AFPM under the dimensional and operating constraints of the existing joint motor, rather than to establish the general superiority of an AFPM over an independently optimized RFPM. The main design specifications and parameters considered in this study are summarized in Table 1.

Table 1. Motor specifications for 260 W target collaborative robot joints.

Dimensions	Radial Flux Motor	Axial Flux Motor	Unit
Rated output	260	↑	W
Rated speed	3500	3500	rpm
Rated current	13	13	Arms
Current limit for EL map	-	18.38	Apk
Voltage limit for EL map	-	48	Vpk
Pole/slot	20/18	-	-
Stator outer/inner diameter	82/54	82/54	mm
Rotor outer/inner diameter	52.2/44.6	82/54	mm
Air gap	0.6	0.6	mm
Stack length	13	-	mm
Axial length	19.6	19.6	mm
Permanent magnet	N42SH	N42SH	-
Stator core	35PN230	Somaloy 700 3P	-

Figure 2 illustrates the structure and magnetic characteristics of the single-stator single-rotor (SSSR) AFPM adopted in this study. Figure 2a shows the main magnetic flux path, while Figure 2b presents the magnetic flux density distribution. The SSSR configuration consists of a single stator and a single rotor, and the magnetic flux generated by the permanent magnets forms a closed magnetic circuit through the air gap, stator, and rotor core. If the rotor core becomes excessively saturated by the permanent magnet flux, the effective flux linkage can be limited, resulting in degradation of the no-load back-EMF and torque performance [4,8].

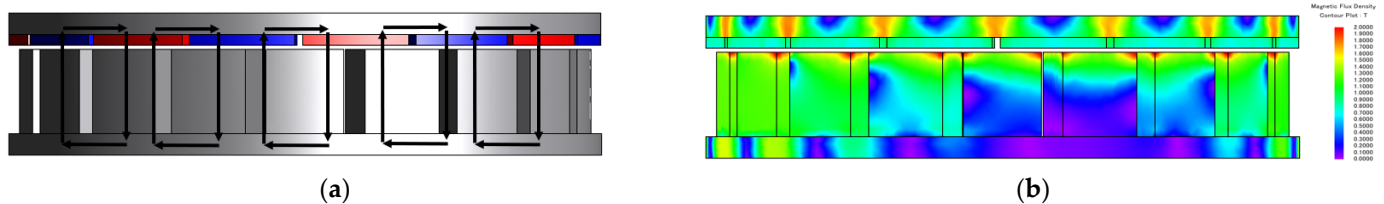


Figure 2. Shape of single stator single rotor AFPM. (a) Magnetic Flux Path (b) Flux Density Distribution.

Therefore, in the detailed geometric design, the no-load back-EMF was not simply maximized; instead, the no-load line-to-line back-EMF and the maximum magnetic flux density under load were simultaneously evaluated for variations in each design variable. Based on these evaluations, a feasible design region was established to achieve the required output performance while preventing excessive magnetic saturation of the core. The value of 1.8 T was selected as an upper design criterion based on the nonlinear magnetization characteristic of the Somaloy 700 3P material used in the FEM model, rather than as a nominal saturation flux density of the material [15]. This criterion was

introduced to avoid excessive operation in the highly nonlinear magnetic region during the detailed design process.

2.2. Pole–Slot Combination Selection Based on Winding Factor and Topological Symmetry

Figure 3 compares the winding factor and the greatest common divisor (GCD) of the numbers of poles and slots for various pole–slot combinations. The pole–slot combination affects the winding factor, spatial harmonics of the magnetomotive force, and torque characteristics; therefore, selecting an appropriate combination is an important step in the preliminary topology design of the motor [12,13].

In this study, the GCD was employed as a preliminary screening index representing the electromagnetic periodicity and symmetry associated with the numbers of poles and slots. Depending on the pole–slot combination, the circumferential periodicity and symmetry of the magnetomotive force distribution may vary, which can influence the spatial distribution of electromagnetic forces. Therefore, pole–slot combinations with an even and relatively large GCD were preferentially considered to identify candidates with sufficient topological symmetry. The GCD was used only as a preliminary topology symmetry index for constructing the initial candidate set.

In addition, because all AFPM candidates considered in this study employ concentrated windings, the winding factor was used as an electromagnetic screening index associated with the fundamental flux linkage, back-EMF, and torque-producing capability [12,13]. Based on the winding factor and GCD, the initial candidate set was identified for subsequent EL map evaluation.

Poles / Slots	12	15	18	21	24	27
8	0.866 / 4					
10	0.933 / 2	0.866 / 5	0.647 / 2			
12			0.866 / 2			
14	0.933 / 2	0.954 / 1	0.902 / 2	0.866 / 7	0.76 / 2	
16	0.866 / 4	0.954 / 1	0.945 / 2	0.89 / 1	0.866 / 8	
18						0.866 / 6
20		0.866 / 5	0.945 / 2	0.953 / 1	0.933 / 4	0.877 / 1
22		0.621 / 1	0.902 / 2	0.953 / 1	0.95 / 2	0.915 / 1
24			0.866 / 6			0.945 / 3
26			0.647 / 1	0.89 / 1	0.95 / 2	0.954 / 1
28	0.5 / 4	0.199 / 1	0.617 / 1	0.866 / 1	0.933 / 4	0.954 / 1

Figure 3. Comparison of winding factor and GCD for different pole–slot combinations. Crossed entries indicate combinations with odd GCD values and asymmetric circumferential periodicity, which may result in less symmetric electromagnetic force distributions.

2.3. Candidate Screening via Multi-Pole EL Map Analysis

Although an initial set of candidates was identified in the previous section based on the winding factor and the symmetry characteristics of the pole–slot combinations, these motor-side indices alone cannot directly determine whether each candidate can generate the target torque required by the actual drive system within the voltage and current

constraints of the inverter. Therefore, in this study, a system-level topology screening method based on an inductance and back-EMF map (EL map) is applied to compare the electromagnetic characteristics of different pole–slot combinations on a common basis and to relate them to the requirements of the actual drive system.

To ensure a consistent preliminary comparison, all pole–slot candidates were evaluated under common initial design conditions before detailed geometric design. This allows the effects of the pole and slot topology to be distinguished from those of topology-specific geometric design. The resulting EL map therefore represents the relative suitability of the candidates under the predefined reference conditions and should not be interpreted as an absolute ranking of fully optimized designs. Because the back-EMF and inductance can vary with detailed geometric parameters, the candidate locations on the EL map may also change if different initial design conditions are adopted.

Under these common conditions, all candidates employed the same concentrated-winding configuration, the same total number of series turns per phase, and the same slot filling factor. Because the available slot area varies with the pole–slot combination, the conductor cross-sectional area was adjusted according to the slot geometry while maintaining the same filling factor. This prevents candidates with a larger number of slots from obtaining an artificial advantage through an increased total number of winding turns or a higher copper utilization ratio. In addition, the same motor envelope, air gap length, permanent magnet and stator core materials, operating speed, target torque, and inverter voltage and current limits were applied to all candidates.

Although the same total number of series turns per phase and slot filling factor were applied to all candidates, the total conductor length and end winding length can vary with the pole–slot combination. Therefore, the phase resistance and copper loss are not necessarily identical among the candidates. In this study, the candidate-dependent total conductor length, including the end winding section, was considered when evaluating the phase resistance and copper loss.

All pole–slot candidate models were analyzed using consistent 3D FEM numerical settings to ensure a fair comparison. Local mesh refinement was applied to the air gap and permanent magnet regions, where relatively large magnetic field gradients occur. The resulting 3D FEM models contained approximately 5.0×10^5 tetrahedral elements. The transient analysis was performed with an angular resolution of five electrical degrees per time step, corresponding to 72 calculation steps per electrical period. The same mesh generation criteria and temporal resolution were maintained for all candidate models. Because detailed AFPM analysis requires computationally intensive 3D FEM, the proposed EL map screening procedure was used to restrict the subsequent detailed parametric design to the selected pole and slot topology. Each pole–slot candidate is represented as a point on the EL map using the synchronous inductance and the fundamental peak value of the no-load phase back-EMF, obtained from 3D FEM under identical initial design conditions, as defined in (1).

$$x_k = (L_k, E_{0,k}) \quad (1)$$

In the EL map, $E_{0,k}$ and L_k are key electromagnetic parameters directly associated with the inverter current and voltage requirements, respectively. The no-load back-EMF is related to the permanent magnet flux linkage and therefore affects the torque constant and the current required to generate the target torque, whereas the inductance determines the reactance voltage according to the electrical angular speed and current. Accordingly, by representing different pole–slot combinations within a common E and L parameter space, the electromagnetic characteristics of each candidate can be compared using

consistent system-level criteria. The relationship between the no-load back-EMF $E_{0,k}$ and the permanent magnet flux linkage $\psi_{f,k}$ is given in (2).

$$E_{0,k} = \omega_{e,k} \psi_{f,k} \quad (2)$$

Here, $\omega_{e,k}$ denotes the electrical angular speed of candidate k . Since the AFPM considered in this study employs a surface-mounted permanent magnet (SPM) configuration, the magnetic saliency is assumed to be small during the initial topology screening stage, and the inductances are approximated as $L_d \approx L_q \approx L_s$. In addition, under $i_d = 0$ control, the electromagnetic torque of candidate k can be expressed as given in (3).

$$T_e = \frac{3}{2} p_k \psi_{f,k} I_k \quad (3)$$

Here, p_k denotes the number of pole pairs of candidate k , and I_k is the peak value of the phase current. In the EL map evaluation, the load torque of the baseline RFPM, 0.737 N·m at the rated speed of 3500 rpm, was defined as the T_{target} . Accordingly, the required current for each pole–slot candidate to generate the target torque is calculated as given in (4).

$$I_{\text{req},k} = \frac{2T_{\text{target}}}{3p_k \psi_{f,k}} \quad (4)$$

By substituting (2) into (4) and applying the relationship $\omega_{e,k} = p_k \omega_m$, the required current can be rearranged as given in (5).

$$I_{\text{req},k} = \frac{2T_{\text{target}} \omega_m}{3E_{0,k}} \quad (5)$$

Here, ω_m denotes the mechanical angular speed. As can be seen from (5), under the same mechanical speed and target torque conditions, a candidate with a higher $E_{0,k}$ requires a lower current to generate the target torque. Therefore, the no-load back-EMF is not only a motor-side performance index but is also directly related to the available current margin and torque-producing capability under the same inverter current constraint.

However, a higher $E_{0,k}$ does not necessarily indicate a more favorable topology. Although an increase in back-EMF reduces the current required to produce the same torque, it may simultaneously increase the voltage required from the inverter. In addition, the reactance voltage determined by the inductance and electrical angular speed affects the voltage requirement of each candidate. Therefore, both the required current and the inverter voltage constraint must be considered when selecting pole and slot topology.

During the initial screening stage, the voltage drop across the stator resistance was neglected by assuming $R_s = 0$, and $i_d = 0$ control was applied. Accordingly, the phase voltage required under the target torque operating condition can be expressed as the vector sum of the no-load back-EMF and the reactance voltage associated with the inductance, as given in (6).

$$V_{\text{req},k} = \sqrt{E_{0,k}^2 + (\omega_{e,k} L_k I_{\text{req},k})^2} \quad (6)$$

Although the no-load back-EMF used for comparison among the candidates is expressed as the line-to-line RMS value, the voltage, current, and back-EMF used in the EL map calculations are all defined based on the peak values of the phase fundamental components. In this study, the peak phase voltage limit of the inverter was set to 48 Vpk. Accordingly, the condition for candidate k to satisfy the inverter voltage constraint is given in (7).

$$V_{req,k} \leq V_{max} = 48Vpk \quad (7)$$

The inverter current constraint is defined as given in (8).

$$I_{req,k} \leq I_{max} \quad (8)$$

In the EL map analysis, the current is also expressed in terms of the peak phase current value. Therefore, the rated current of 13 Arms listed in Table 1 was converted to 18.38 Apk for application as the current limit in the EL map. To evaluate the relative torque-producing capability among the candidates satisfying both the voltage and current constraints, the current margin of each candidate is defined as given in (9).

$$\Delta I_k = I_{max} - I_{req,k} \quad (9)$$

A larger current margin defined in (9) indicates that additional current capacity remains after generating the target torque. Therefore, a candidate with a larger current margin can be considered to have a relatively greater potential for torque expansion under the same inverter current constraint. Accordingly, in this study, the current margin was used as the primary ranking criterion among the candidates satisfying both the voltage and current constraints.

Although a higher $E0,k$ is advantageous for reducing the required current, as can be seen from (6), the back-EMF itself increases the required voltage. Therefore, an excessively high $E0,k$ may be disadvantageous with respect to the inverter voltage constraint. In addition, even for candidates with the same back-EMF, the reactance voltage may differ due to variations in the electrical angular speed $\omega e,k$ associated with the number of poles and the inductance Lk . Therefore, rather than simply selecting the candidate with the highest back-EMF, the pole–slot combinations were first evaluated based on whether the required current and voltage at the target torque operating condition simultaneously satisfied the inverter constraints.

Accordingly, each candidate, including k , $I_{req,k}$ and $V_{req,k}$, was calculated from $(Lk, E0,k)$, and the candidates satisfying the condition in (10) were classified as feasible candidates.

$$A = \left\{ k \mid I_{req,k} \leq I_{max}, V_{req,k} \leq V_{max} \right\} \quad (10)$$

Accordingly, in the proposed pole and slot selection procedure, the system-level feasibility of each candidate with respect to the inverter voltage and current constraints was first evaluated using (10). Candidates satisfying both the current and voltage constraints were classified as feasible, those exceeding the current limit were classified as current limited, and those satisfying the current limit but exceeding the voltage limit were classified as voltage limited. Among the feasible candidates, the current margin defined in (9) was used as the primary ranking criterion, while the total harmonic distortion (THD) of the no-load back-EMF was additionally considered as a waveform quality index when selecting the final pole and slot topology for detailed design.

In this study, the EL map serves as a system-level evaluation framework that directly links $E0,k$ and Lk , which represent the electromagnetic characteristics of each pole–slot combination, to the target torque and the inverter voltage and current constraints. In other words, rather than independently comparing individual parameters such as the winding factor, back-EMF, or inductance, the main purpose of the EL map is to evaluate different pole and slot topologies within a common E and L parameter space and to compare their suitability for the actual drive-system conditions using consistent criteria.

3. Results and Discussion

3.1. EL Map Screening Results

Figure 4 shows the pole–slot candidates on the EL map together with the inverter current and voltage constraints, while Figure 5 presents an enlarged view of the candidate region. The background color represents the required phase peak current $I_{\text{req},k}$ for generating the target torque. The horizontal I_{max} boundary corresponds to the inverter current limit of 18.38 Apk. Because the electrical angular speed $\omega_{e,k} = p_k \omega_m$ varies with the number of pole pairs, the voltage limit boundary is different for each pole number. Accordingly, multiple dashed curves are shown for V_{max} . For each pole number, the upper and lower branches represent the two solutions of the voltage limit equation under the target torque condition. Figure 5 enlarges the region in which the pole–slot candidates are concentrated to improve the readability of their relative positions.

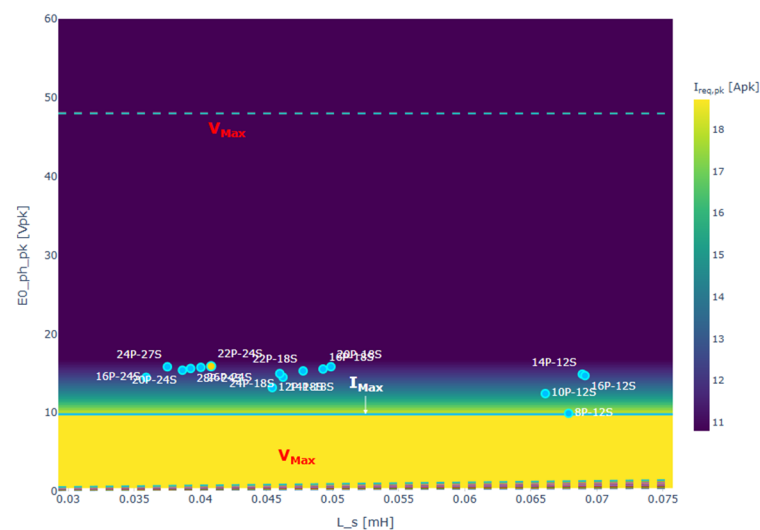


Figure 4. EL map of all pole–slot candidates with inverter current and pole-dependent voltage constraints.

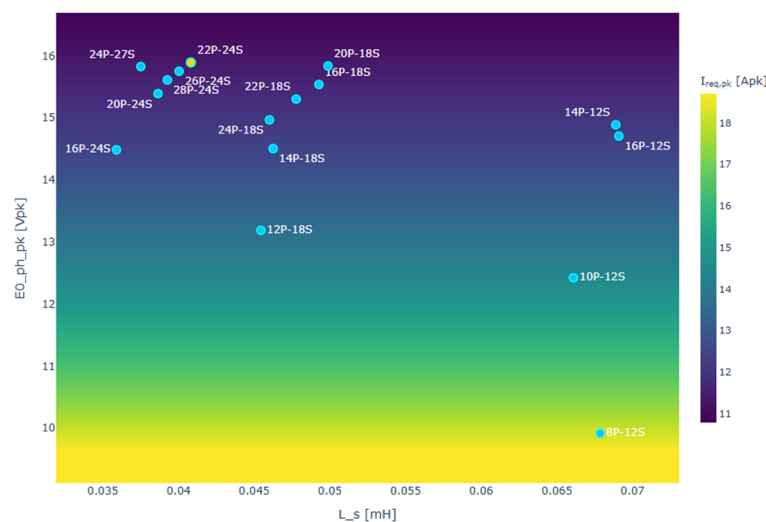


Figure 5. Enlarged EL map of the pole–slot candidate region.

Based on the EL map evaluation, the 22-pole/24-slot combination satisfied the inverter voltage and current constraints and provided a relatively large current margin

among the feasible candidates. The no-load back-EMF waveform characteristics shown in Figure 6 were additionally considered as a secondary selection criterion. Based on the combined evaluation of system-level feasibility, current margin, and back-EMF waveform quality, the 22-pole/24-slot combination was selected for the subsequent detailed 3D FEM design.

Poles / Slots	12	15	18	21	24	27
8	12.23 Vrms / 14.46%					
10	15.22 Vrms / 3 %					
12			16.20 Vrms / 11.78 %			
14	18.24 Vrms / 2.95 %		17.77 Vrms / 1.85 %			
16	17.91 Vrms / 11.08 %		19.12 Vrms / 1.85 %		17.77 Vrms / 11.01 %	
18						
20			19.36 Vrms / 5.96 %		19.02 Vrms / 2.92 %	
22			18.73 Vrms / 2.55 %		19.45 Vrms / 2.9 %	
24			18.29 Vrms / 6.19 %			19.36 Vrms / 3.93 %
26					19.28 Vrms / 0.64 %	
28					19.14 Vrms / 0.66 %	

Figure 6. Distribution of no-load line voltage and THD values.

3.2. Detailed 3D FEM Parametric Design

Figure 7a illustrates the design variables of the AFPM. Considering manufacturability, the stator core length (L_{Stator}) was fixed at 3 mm, which is the minimum feasible length for fabrication. The detailed design was conducted as a sequential discrete parametric study rather than as a global multi-variable optimization. Each data point shown in Figure 7b–e corresponds to an individual 3D FEM analysis at a discrete value of the corresponding design variable. At each design step, the selected parameter value was fixed and carried forward to the subsequent step. Therefore, the sequential procedure is not intended to identify a global optimum over the entire multidimensional design space, but rather to determine a suitable design within the investigated parameter ranges while satisfying the predefined dimensional and magnetic saturation constraints. At each step, one design variable was varied over a predefined range while the variables selected in the preceding steps were fixed. Step 1 determines the number of turns. Step 2 determines the permanent magnet length. Step 3 determines the permanent magnet gap, and Step 4 determines the tooth spacing. Step 1 (Number of Turns Selection): The number of turns is an important design variable that directly affects the no-load back-EMF and torque constant of the motor. Under the fixed stator geometry and winding fill factor constraint, the number of turns was varied and the corresponding no-load line-to-line back-EMF was evaluated using 3D FEM. As shown in Figure 7b, the back-EMF increases with the number of turns and then tends to saturate. Therefore, the number of turns was selected considering the required back-EMF, available slot area, and winding manufacturability. Step 2 (Permanent Magnet Length Selection): Under the established rotor length constraint, L_{Magnet} was varied to simultaneously evaluate the no-load back-EMF and the magnetic flux density distribution. The rotor core saturation was verified according to the magnet length, thereby determining a selected permanent magnet length that satisfies the predefined

magnetic saturation condition. Step 3 (Permanent Magnet Spacing Selection): The gap between adjacent permanent magnets affects the leakage flux and magnetic saturation of the core. Therefore, the permanent magnet spacing (G_{Magnet}) was selected by evaluating the no-load back-EMF while satisfying the predefined magnetic saturation constraint over the investigated range. Step 4 (Tooth Spacing Selection): The slot area, which dictates the feasible number of winding turns, is determined by the tooth spacing and tooth length. In a shoeless structure, excessively narrow tooth spacing can lead to severe magnetic flux density saturation within the teeth. Therefore, the selected tooth spacing (T_{Teeth}) was determined by analyzing the no-load back-EMF while complying with the saturation constraint across varying tooth spacings. The corresponding parametric analysis of the no-load line-to-line voltage for these AFPM design variables is summarized in Figure 7b–e.

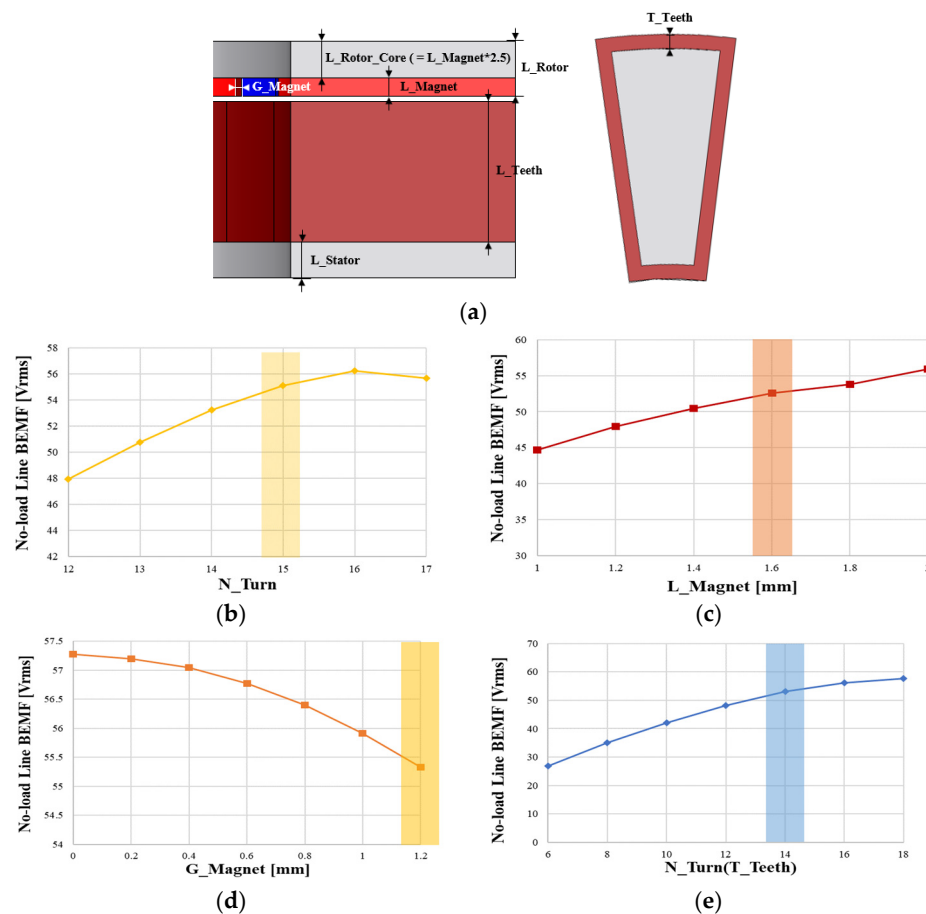


Figure 7. (a) Design variables of the AFPM and parametric analysis of the no-load line-to-line voltage with respect to (b) number of turns, (c) permanent magnet length L_{Magnet} , (d) permanent magnet spacing G_{Magnet} , and (e) tooth spacing T_{Teeth} . The colored vertical bars indicate the parameter values selected at each sequential design step.

The AFPM performance reported in Table 2 corresponds to the final design obtained using the parameter values highlighted by the vertical bars in Figure 7b–e.

Table 2. Comparison of the baseline RFPM and final AFPM.

Dimensions	Radial Flux Motor	Axial Flux Motor	Unit
No-load line voltage	29.99	40.19	Vrms
No-load line THD	1.86	0.65	%
Load torque	0.737	0.907	N·m

Copper loss	12.77	7.58	W
Core loss	6.17	29.3	W
Power	263.60	332.4	W
Efficiency	93.17	89.23	%

The two machines exhibit different loss characteristics. The AFPM shows a lower copper loss than the baseline RFPM, whereas its core loss is substantially higher. The relatively small difference in pole number and electrical frequency alone cannot account for the observed core loss difference. As shown in Table 2, the two machines employ different stator core materials and also have fundamentally different core geometries and magnetic flux paths, resulting in different magnetic flux density distributions. The load magnetic flux density distribution of the final AFPM design is shown in Figure 8. Therefore, the difference in core loss is attributed to the combined effects of the core material characteristics, magnetic loading, and core geometry rather than to the pole number alone. Further reduction of the AFPM core loss remains an important subject for subsequent design refinement.

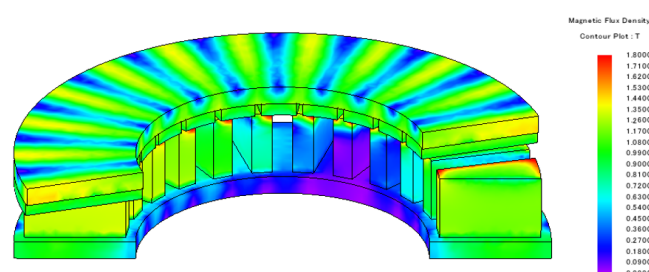


Figure 8. Load magnetic flux density of the final design.

4. Conclusions

This study proposed an inverter constraint-based EL map screening method for selecting a pole–slot combination of an AFPM for collaborative robot joints. The pole–slot candidates were evaluated based on the inverter voltage and current constraints, current margin, and no-load back-EMF waveform characteristics, and the 22-pole/24-slot combination was selected for detailed 3D FEM design. Under the predefined dimensional and operating constraints, the final AFPM achieved a load torque of 0.907 N·m and an output power of 332.4 W, compared with 0.737 N·m and 263.6 W for the baseline RFPM. These results correspond to improvements of 23.1% in torque density and 26.1% in power density. However, the increased core loss of the AFPM remains an important issue for further design improvement. The present study is based on 3D FEM analysis, and experimental validation of the selected AFPM remains as future work.

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