

Research Paper

Optimized Flux-Weakening Strategy in Field-Oriented Control for High-Speed IPMSM Drives

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Abstract

Unlocking the dynamic performance of electric vehicles is often limited by the voltage constraints of the battery system. This study proposes an optimized control strategy for Interior Permanent Magnet Synchronous Motors based on an analytical formulation of direct-axis current trajectories to maximize speed extension while maintaining torque smoothness under strict battery voltage constraints, utilizing parameters characteristic of commercial C-segment electric vehicles (e.g., VinFast VF e34). Through a comprehensive simulation framework, the research investigates a coordinated Field-Oriented Control scheme integrated with a Flux-Weakening strategy through direct-axis current adjustment to reconcile the conflict between high-speed operation (up to 6 times the base speed of 100 rad/s) and power quality. The analysis identifies a critical operating point at a direct-axis current of -25 A, which effectively prevents voltage saturation while maintaining torque smoothness. The results demonstrate that, when evaluated against the baseline Field-Oriented Control without flux-weakening at 600 rad/s, this specific trajectory significantly reduces torque ripple by 8.6 Nm and suppresses Total Harmonic Distortion to a negligible 0.19%. This combined mitigation contributes to the high-speed operating capability by preventing system oscillations and preserving linear voltage modulation at this upper speed limit. These findings provide a validated guideline for enhancing powertrain stability and mechanical lifespan in modern electric mobility.

Keywords: Electric vehicle; IPMSM; Field-Oriented Control; Flux-weakening

1. Introduction

In recent years, the global automotive industry has witnessed a rapid shift toward sustainable mobility, with Electric Vehicles (EVs) emerging as a key solution to reduce carbon emissions [1]. Within this landscape, the Interior Permanent Magnet Synchronous Motor (IPMSM) has established itself as the dominant choice for EVs due to its superior power density, high efficiency, and wide operating speed range [2]. Nevertheless, a challenge in EV powertrains is the rigid voltage constraint imposed by the battery pack (DC bus). As rotational speed increases, the induced back-electromotive force (Back-EMF) rises proportionally and eventually exceeds the

maximum output voltage of the inverter. If left unaddressed, this voltage saturation phenomenon creates a sharp decline in torque production and current controllability, thereby limiting the high-speed dynamic performance of EVs [3].

To mitigate this limitation, the implementation of Flux-Weakening (FW) control within the Field-Oriented Control (FOC) framework has become the standard approach in modern EV drives. As established in the foundational work by Jahns [4], the core principle involves injecting a negative direct-axis current to counteract the permanent magnet flux, thereby lowering the effective back-EMF and widening the operating speed range beyond the base speed. While various advanced



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algorithms, such as Model Predictive Control [5] or Sliding Mode Observers [6] have been proposed to enhance this process, the feedback-based FW technique remains favored in industrial applications due to its robustness and simplicity [7].

However, a critical gap remains in the comprehensive optimization of this strategy. Most existing studies primarily focus on maximizing the speed range, often neglecting the degradation of power quality that occurs during deep FW operation [6], [7]. Excessive current injection to force high-speed operation can lead to severe torque ripples and harmonic distortion, which shorten the mechanical lifespan of the drivetrain. Consequently, this research seeks to bridge the gap between theoretical speed extension and mechanical stability. The study utilizes validated motor parameters typical of a commercial C-segment EV, specifically modeled after the VinFast VF e34, a prominent electric crossover developed in Vietnam. By analyzing this specific traction motor topology, we determine the optimal equilibrium between speed expansion and torque smoothness, ensuring robust performance that complies with international standards [8], [9]. The overall research procedure, ranging from mathematical modeling to strategy verification, is visually summarized in Figure 1.

2. Methods

2.1. Mathematical Modeling of IPMSM

To formulate the control strategy, the mathematical model of the VinFast VF e34 traction motor is described in the synchronous rotating reference frame ($d-q$). By using Clarke and Park transformations, the three-phase AC variables are converted into DC components to simplify the control design [10]. This modeling approach is built upon the foundational driver circuit design and control techniques previously investigated in [11], [12]. The electrical dynamics of the motor are defined by the $d-q$ axis voltage balance equations, as given by Eq. (1) and Eq. (2):

$$v_d = R_s i_d + L_d \frac{di_d}{dt} - \omega_r L_q i_q \quad (1)$$

$$v_q = R_s i_q + L_q \frac{di_q}{dt} + \omega_r L_d i_d + \omega_r \psi_f \quad (2)$$

where, v_d, v_q and i_d, i_q are the stator voltages and currents; R_s is the stator resistance; L_d, L_q are the inductances; and ω_r is the electrical angular velocity, ψ_f is the magnetic flux. Since the VF e34 employs an IPMSM topology ($L_d < L_q$), the electromagnetic torque (T_e) [13], [14] includes both permanent magnet alignment torque and reluctance torque, as expressed in Eq. (3):

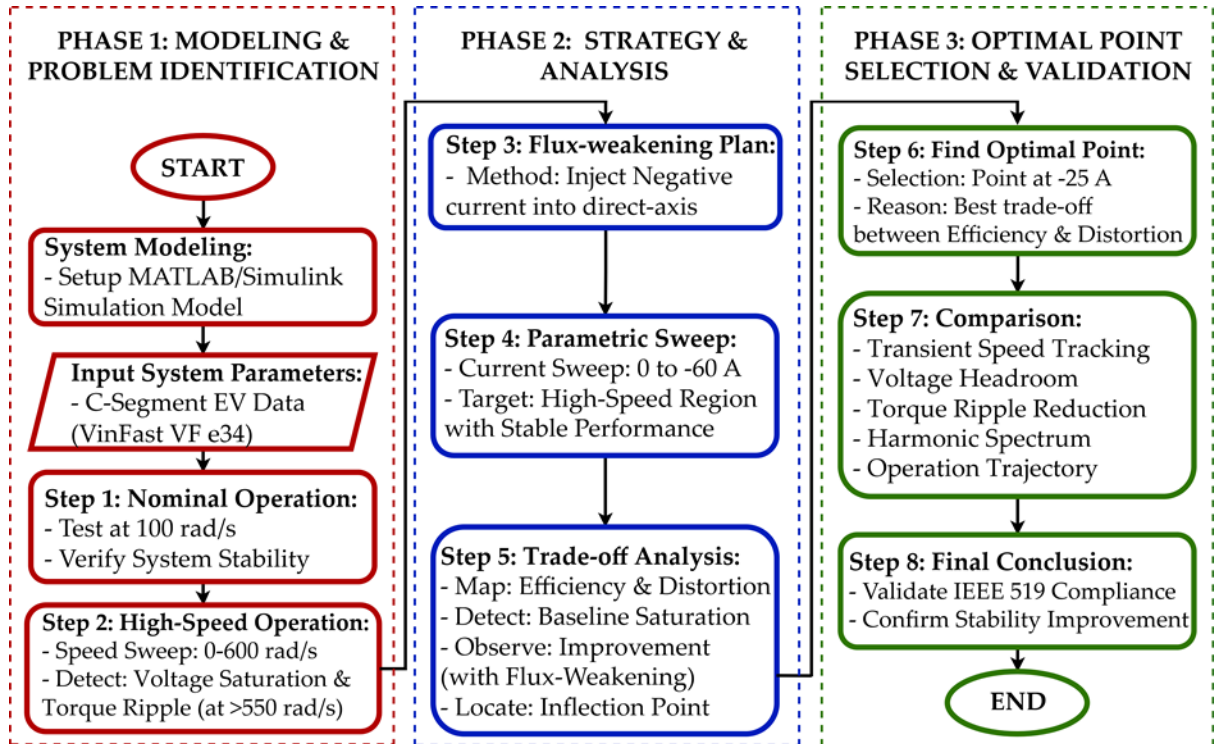


Figure 1. Overall flowchart of the proposed research framework

$$T_e = \frac{3}{2}p[\psi_f i_q + (L_d - L_q)i_d i_q] \quad (3)$$

The electromechanical coupling is defined by the Motion Eq. (4):

$$\frac{d\omega_m}{dt} = \frac{1}{J} (T_e - T_L - B\omega_m) \quad (4)$$

where, ω_m is the mechanical rotor speed, J is the moment of inertia, B is the viscous friction coefficient, and T_L is the load torque. These general equations form the constraints for the high-speed FW control.

2.2. Field-Oriented Control Technique

2.2.1. Field-oriented control principle

The FOC technique aims to decouple Permanent Magnet Synchronous Motor (PMSM) stator current into two orthogonal components, allowing independent control of magnetic flux

and electromagnetic torque. This approach mimics the excellent control characteristics of a separately excited DC motor [15], [16]. The decoupled control structure of a DC motor is illustrated in **Figure 2**, which is adapted from the concept presented in [15].

In this structure, the field current (I_f) and armature current (I_a) are mechanically orthogonal (90°). I_f establishes ψ_f on the d -axis, while I_a flows along the q -axis. The resulting T_e is generated by the interaction of ψ_f and I_a . This decoupling ensures torque control via I_a without disturbing the flux. **Figure 3** depicts how this principle is applied to PMSM, based on the representation in [15].

FOC emulates DC Motor characteristics in a IPMSM by aligning the rotating d -axis with the ψ_f at angle θ . By projecting the $a - b - c$ stationary frame onto the $d - q$ synchronous rotating frame,

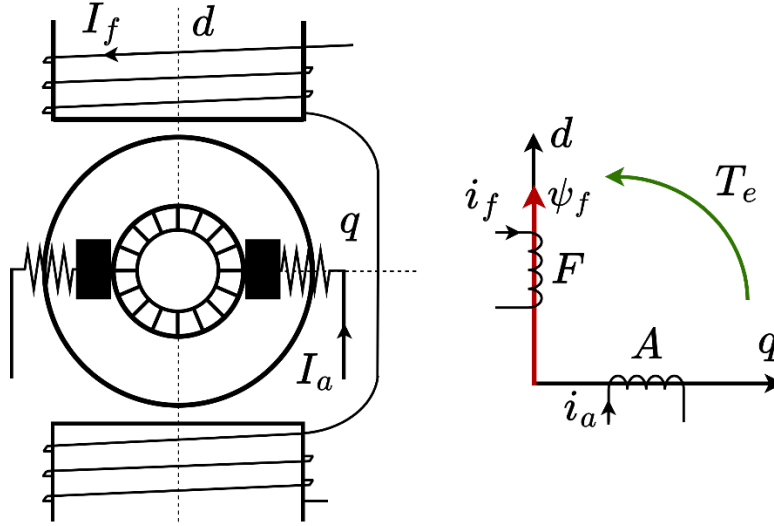


Figure 2. Brush DC motor decoupled control

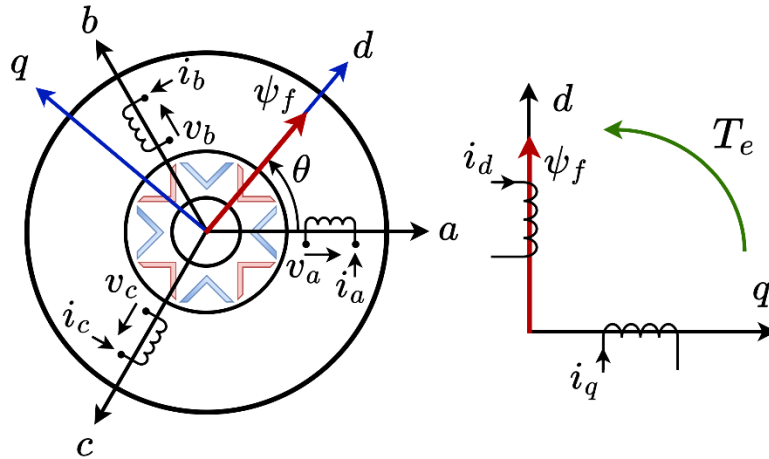


Figure 3. FOC principle applied to PMSM

the system becomes a decoupled equivalent model where i_d controls flux (same as I_f), while i_q controls the torque (same as I_a). Consequently, the generated torque (T_e) follows the same vector interaction, separating flux and torque dynamics as mirrored in the DC motor structure.

2.2.2. Baseline control strategy

In the preliminary design stage, a Baseline Control Strategy is implemented by setting d-axis current to zero. This approach is based on the ideal Surface-mounted Permanent Magnet Synchronous Motor (SPMSM) characteristics, where the reluctance torque component is naturally absent. Consequently, the reluctance term in Equation (3) is eliminated. To simplify the control architecture and minimize stator resistive losses, the reference current is defined as [15] by Eq. (5):

$$i_d^* = 0 \quad (5)$$

Substituting the control condition in Eq. (5) into Eq. (3), the electromagnetic torque expression can be linearized and simplified to Eq. (6):

$$T_e = \left(\frac{3}{2}p\psi_f\right)i_q = K_T \cdot i_q \quad (6)$$

where, K_T is the torque constant, indicates that the electromagnetic torque is directly proportional to the q -axis current, making a dynamic response similar to a DC motor [17].

However, commercial traction motors for EVs, specifically the VF e34 motor investigated in this work, are typically IPMSM topologies exhibiting distinct magnetic saliency ($L_d < L_q$). While the zero d -axis current approach (5) offers simplicity, it fails to harness the full torque potential. To address this problem, Section 3.4 introduces a negative current injection strategy with $i_d^* < 0$. This enables FW capabilities and exploits the inequality ($L_d < L_q$) to generate additional torque, thus maximizing energy efficiency.

2.2.3. Field-oriented control scheme

The overall control architecture, as shown in Figure 4, adapts the standard FOC structure presented in [15], [16], [18] into a dedicated coordinated scheme. Specifically, the system features critical modifications that include direct-axis current adjustments via the FW algorithm to systematically manage voltage boundaries, alongside the implementation of the Sinusoidal

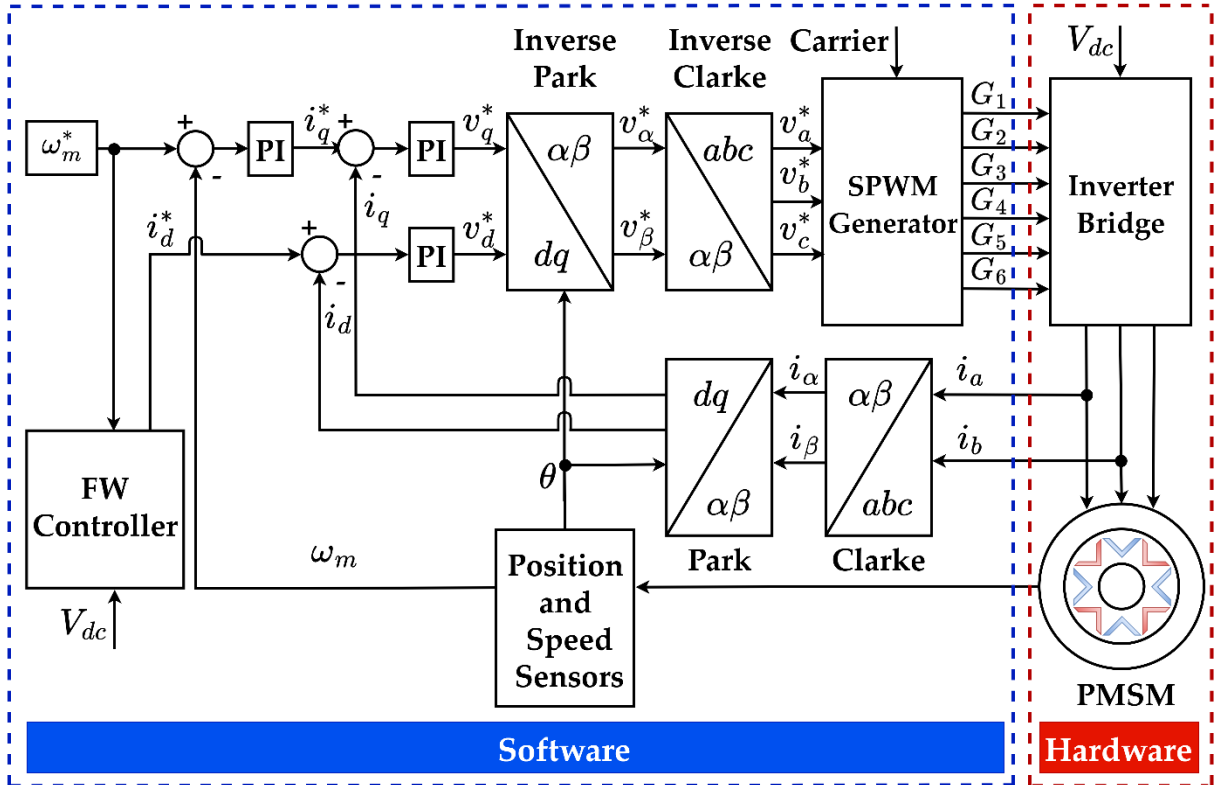


Figure 4. Block diagram of the PMSM FOC scheme with SPWM technique

Pulse Width Modulation (SPWM) technique tailored for the VF e34 drive system.

The control strategy uses a cascaded loop structure, beginning with the outer speed loop where the measured rotor speed (ω_m) and the reference speed (ω_m^*) are compared [17]. The resulting error is processed by a PI controller to create a quadrature current reference (i_q^*), representing the torque command. Then, the measured phase currents (i_a, i_b) are converted into the rotating reference frame (i_d, i_q) through Clarke and Park transformations [15]. These currents are compared with their respective references (i_d^*, i_q^*), enable the PI controllers to find the errors and produce the reference voltages in the $d-q$ rotating frame (v_d^*, v_q^*) [16].

To execute the SPWM method, the reference voltages are changed back to the stationary frame (v_a^*, v_b^*, v_c^*) by Inverse Park and Clarke transformations. These sinusoidal reference signals are compared with a high-frequency triangular carrier to produce six gate signals (G_1 to G_6) for the Inverter Bridge, modulating the DC bus voltage (V_{dc}) and powering the PMSM. Finally, a Position & Speed Sensor brings necessary feedback information (θ, ω_m) to close the control loop.

2.3. Controller Design

2.3.1. Proportional-Integral controller

Proportional-Integral (PI) controllers are used for the precise regulation of rotor speed and stator currents. Although conventional Proportional-Integral-Derivative controllers produce faster transient responses, they are unsuitable due to the derivative term. It is sensitive to high-frequency noise from inverter and sensors. Therefore, the PI controller is preferred for stability and simplicity. This process is finding the error signal ($e(t)$), which is the difference between the reference value ($r(t)$) and the measured feedback value ($y(t)$) as Eq. (7):

$$e(t) = r(t) - y(t) \quad (7)$$

The control output $u(t)$ in the continuous time domain [19] is expressed in Eq. (8):

$$u(t) = K_p e(t) + K_i \int_0^t e(\tau) d\tau \quad (8)$$

Where K_p and K_i are the proportional and integral gains, while t is the instantaneous time and τ is the variable of integration.

2.3.2. Sinusoidal Pulse Width Modulation technique

The Sinusoidal Pulse Width Modulation (SPWM) technique is applied to generate gate signals for the three-phase inverter. This method modulates the output voltage to approximate a sinusoidal waveform while displacing harmonic components to higher frequencies for better filtering [20]. The principle of SPWM generation is detailed in Figure 5.

SPWM generation can be achieved by comparing a low-frequency sinusoidal reference ($v_{control}$) against a high-frequency triangular carrier ($v_{carrier}$) as in [12]. The switching logic for every phase leg x (with $x \in \{a, b, c\}$) is determined by Eq. (9):

$$S_x = \begin{cases} 1 & \text{if } v_x^* \geq v_{carrier} \\ 0 & \text{if } v_x^* < v_{carrier} \end{cases} \quad (9)$$

Here, S_x stands for the switching state of the upper Insulated Gate Bipolar Transistor (IGBT) in phase x (1 for ON, 0 for OFF), whereas the lower IGBT operates in a complementary manner.

A key parameter in SPWM is the Amplitude Modulation Index (m_a), which defines the relation-

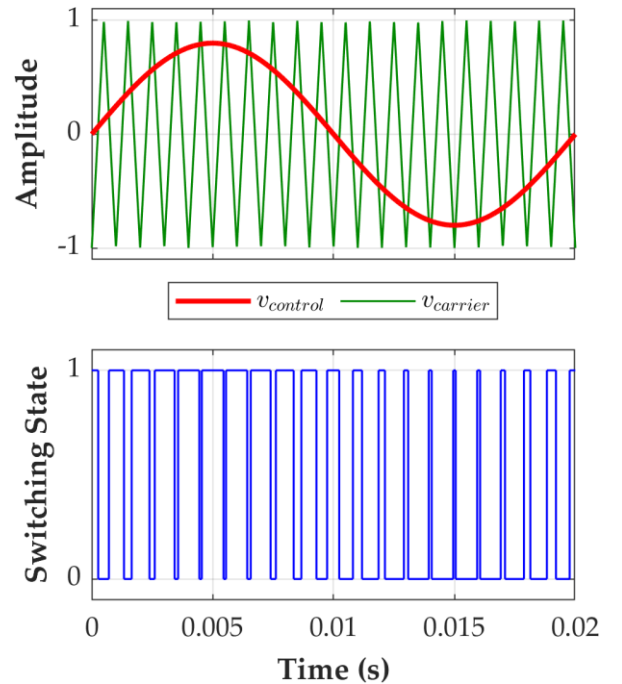


Figure 5. Principle of sinusoidal pulse width modulation

ship between the reference signal and DC bus utilization. It is expressed as the ratio of the peak amplitudes of the control and carrier signals [21], expressed in Eq. (10):

$$m_a = \frac{V_{control}}{V_{carrier}} \quad (10)$$

2.4. Three-Phase Inverter Circuit Schematic

The three-phase Voltage Source Inverter (VSI) connected to the PMSM includes a DC source (V_{dc}), a DC-link capacitor (C_{dc}), and a six-switch inverter bridge (S_1 to S_6) [22]. Antiparallel diodes are included with the IGBTs to handle inductive currents. On the motor side, the PMSM is modeled by its per-phase equivalent circuit, comprising the stator resistance (R_s), synchronous inductance (L_s), and the induced back-electromotive forces (e_a, e_b, e_c), as illustrated in Figure 6.

To avoid short circuits across the DC bus, both switches in the same leg (e.g., S_1 and S_4 for phase A) must operate complementarily. Accordingly, the state of each phase leg x (with $x \in \{a, b, c\}$) is represented by the binary switching function S_x [22] using Eq. (11):

$$S_x = \begin{cases} 1 & \text{if upper switch is ON, lower switch is OFF} \\ 0 & \text{if upper switch is OFF, lower switch is ON} \end{cases} \quad (11)$$

For a balanced Y-connected stator winding with an isolated neutral point, the relationship between the applied phase voltages (v_{an}, v_{bn}, v_{cn}), the DC bus voltage, and the switching states can be explained in matrix form as [23] using Eq. (12):

$$\begin{bmatrix} v_{an} \\ v_{bn} \\ v_{cn} \end{bmatrix} = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} \quad (12)$$

Eq. (12) demonstrates the direct control of output phase voltages through the switching vector $S = [S_a, S_b, S_c]^T$ by the SPWM logic. Furthermore, the DC-link capacitor C_{dc} is important to filter out high-frequency switching harmonics and stabilize the DC bus voltage.

2.5. Power and Efficiency Calculation

Based on the fundamental dynamic model of PMSM [24], the input electrical power (P_{in}) is found by using the three-phase phase voltages and currents, whereas the output mechanical power (P_{out}) is determined by the load torque (T_L) and the rotor mechanical speed, calculated via Eq. (13) and (14):

$$P_{in}(t) = v_a i_a + v_b i_b + v_c i_c \quad (13)$$

$$P_{out}(t) = T_L \cdot \omega_m \quad (14)$$

The average efficiency (η) [25] is computed based on filtered power signals, effectively removing high-frequency switching noise in the inverter operation using Eq. (15):

$$\eta = \frac{P_{out,avg}}{P_{in,avg}} \times 100 \quad (15)$$

2.6. Flux-Weakening and Harmonic Mitigation Strategy

The operational range of the PMSM drive is bounded by the inverter's current rating and the available DC bus voltage. As the rotor accelerates, the Back-EMF rises proportionally. FW strategy is implemented by injecting negative d -axis current ($i_d < 0$) to counteract the effective rotor flux. Neglecting the stator resistance at high speeds, the system is constrained by the voltage limit ellipse

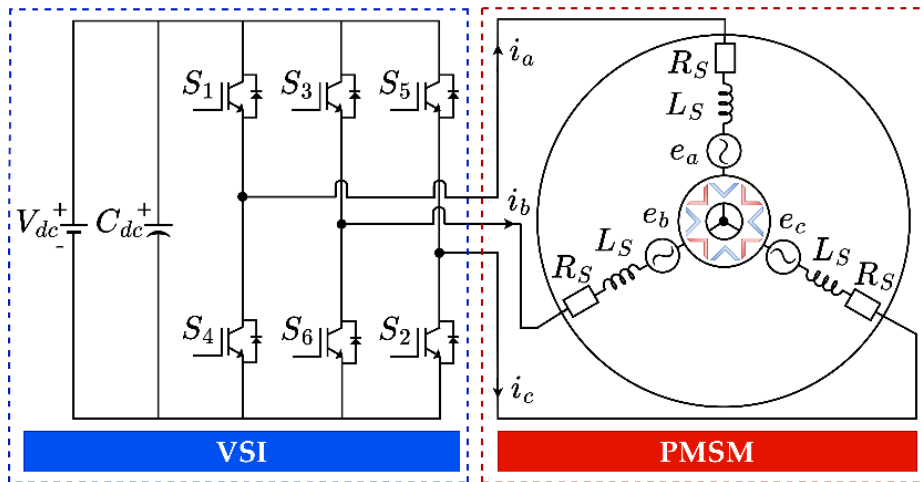


Figure 6. Three-phase voltage source inverter to PMSM diagram

and the current limit circle [4], as defined in Eq. (16) and (17), respectively:

$$V_s \approx \omega_r \sqrt{(L_d i_d + \psi_f)^2 + (L_q i_q)^2} \leq V_{max} = \frac{V_{dc}}{2} \quad (16)$$

$$I_s = \sqrt{i_d^2 + i_q^2} \leq i_{max} \quad (17)$$

Where V_s is the stator voltage magnitude and I_s is the current magnitude. However, operating near the voltage boundary or within the overmodulation zone of the SPWM technique inevitably introduces low-order harmonics (5th, 7th, 11th, 13th) into the currents [7]. To assess the strategy in preventing these distortions, we used Fast Fourier Transform (FFT) to analyze the Total Harmonic Distortion (THD) index [26] using Eq. (18):

$$THD(\%) = \frac{\sqrt{\sum_{n=2}^{\infty} I_n^2}}{I_1} \times 100 \quad (18)$$

Here, I_1 is the fundamental current amplitude and I_n is the amplitude of the n -th harmonic order.

3. Result and Discussion

3.1. Simulation Parameters

The proposed FOC algorithm is validated through simulation model in MATLAB/Simulink environment [27], using VinFast VF e34 as a representative case study. The macroscopic data are derived from the official datasheet [28] and characterized according to the UN ECE Regulation No. 85 test procedure [29]. Meanwhile, proprietary internal parameters are extrapolated from the Nissan Leaf traction motor [30] based on comparable IPMSM topology. The inverter switching frequency is standardized at 10 kHz to reflect typical powertrain standards. Key electromagnetic parameters are listed in Table 1.

3.2. Model Verification under Nominal Conditions

Prior to investigating high-speed instability, the fundamental integrity of the FOC-SPWM algorithm is evaluated under a nominal condition with a reference speed $\omega_m^* = 100$ rad/s, which serves as the base speed in this study. The investigation further extends the operating range up to 600 rad/s (6 times the base speed) to evaluate the high-speed performance and torque ripple characteristics later on in Section 3.3. The steady-

state system response, illustrated in Figure 7, demonstrates that the rotor speed accurately tracks the reference trajectory with high fidelity. Correspondingly, the electromagnetic torque stabilizes at approximately 20 Nm, maintained by a quadrature current of around 38 A. This performance confirms the effectiveness of the PI controllers in both inner and outer loops.

The modulation signals and gating pulse generation are detailed in Figure 8. Consequently, the resulting three-phase stator current waveforms are presented in Figure 9. The precise interaction between the triangular carrier and the sinusoidal control signal, generates proper gating pulses. As a result, the stator currents exhibit a balanced sinusoidal profile with a magnitude of about 37.5 A. Despite the presence of minor high-frequency ripples due to the switching nature, the overall waveform quality ensures smooth torque generation, establishing a reliable baseline for the subsequent FW analysis.

3.3. Problem Statement of Performance Degradation at High Speed

While the system performs robustly at nominal speeds, substantial performance degradation occurs when testing at higher speeds. To investigate this, the Baseline Control Strategy

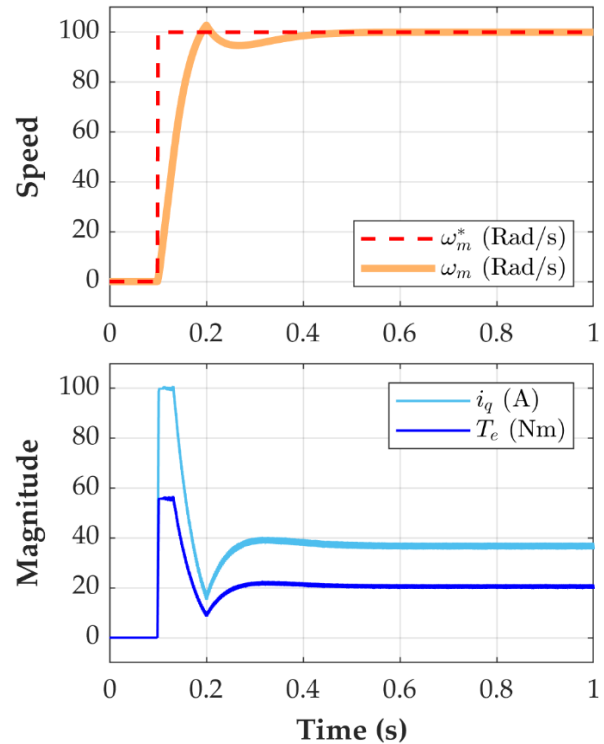
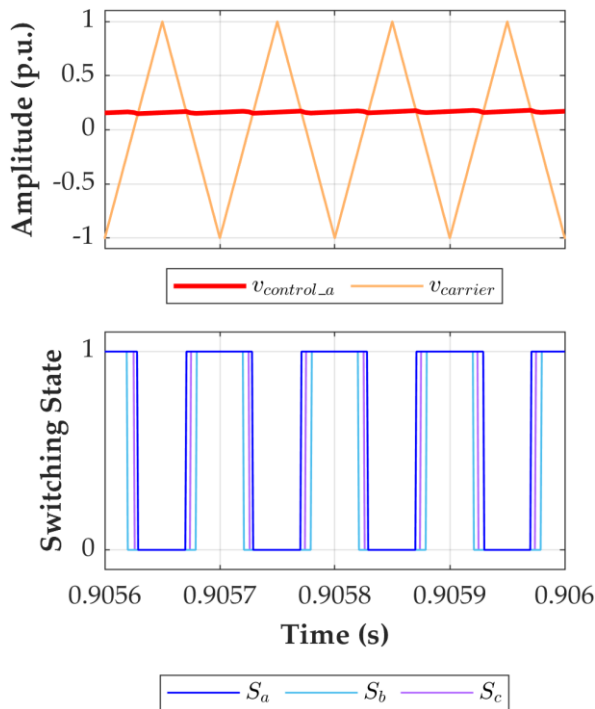
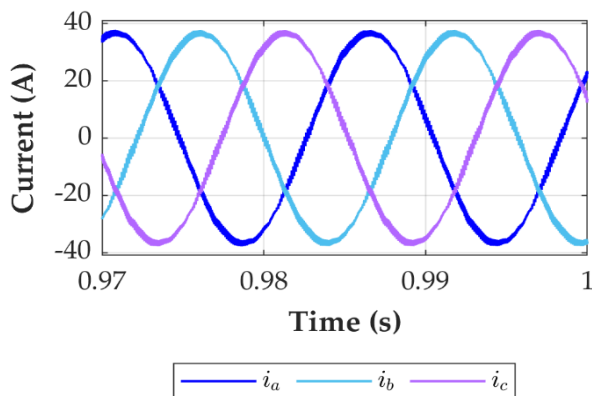


Figure 7. Steady-state speed and torque response at 100 rad/s

Table 1. Specification and electromagnetic parameters of the traction motor

Parameter	Symbol	Value	Unit
Rated Peak Power	P_{rated}	110	kW
Max Torque	T_{max}	242	Nm
DC Bus Voltage	V_{dc}	400	V
Pole Pairs	p	4	pairs
Stator Resistance	R_s	0.026	Ω
Direct-axis stator inductance	L_d	255	μH
Quadrature-axis stator inductance	L_q	750	μH
PM Flux Linkage	ψ_f	0.093	Wb
Moment of Inertia	J	0.035	$\text{kg} \cdot \text{m}^2$
Viscous Friction	B	0.005	$\text{N} \cdot \text{m} \cdot \text{s} / \text{rad}$

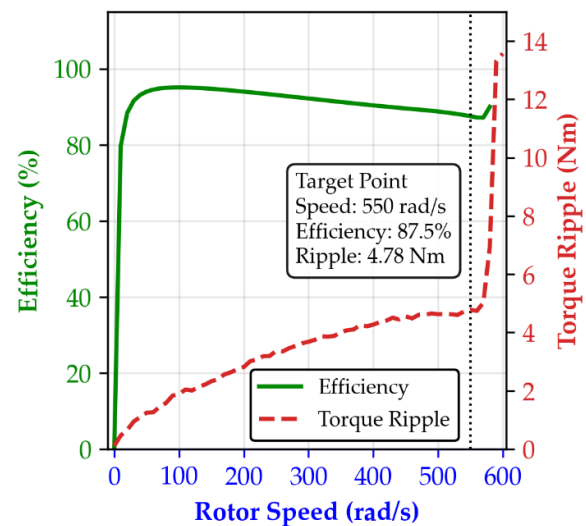
**Figure 8.** SPWM modulation signals and corresponding gate pulses at 100 rad/s**Figure 9.** Three-phase stator current waveforms at 100 rad/s

($i_d^* = 0$) was tested across a speed range from 0 to 600 rad/s, with the efficiency and torque ripple characteristics plotted in **Figure 10**.

The results reveal a critical stability threshold. Up to 500 rad/s, the system maintains high performance. However, a distinct deterioration starts at 550 rad/s, where the torque ripple intensifies to 4.78 Nm, although the efficiency remains acceptable at 87.5%. This condition marks the onset of voltage saturation, where the inverter runs out of linear control margin. Beyond this point, the instability escalates exponentially; specifically at 600 rad/s, the ripple spikes uncontrollably to nearly 14 Nm, posing a severe risk of mechanical fatigue and passenger discomfort. Consequently, 550 rad/s is identified as the critical boundary condition and is selected as the primary target for optimization to restore stability before catastrophic loss of control occurs.

3.4. Optimization Strategy Identification

In response to the high-speed instability identified in Section 3.3, a parametric sweep of the flux-weakening current (i_d^*) from 0 A to -60 A was conducted. The critical trade-off between

**Figure 10.** Diagram of efficiency and torque ripple versus rotor speed ($i_d^* = 0$)

energy efficiency and current total harmonic distortion, is illustrated in Figure 11.

The analysis indicates a distinct inflection point at $i_d^* = -25$ A. Under the baseline condition ($i_d^* = 0$ A), the system suffers from distortion due to voltage saturation, resulting in a THD of approximately 2.56%. By injecting $i_d^* = -25$ A, the voltage vector is pulled back into the linear modulation region, causing the THD to be suppressed drastically to 0.19%, which is crucial for high-speed operation as it prevents voltage saturation at the maximum tested speed. Although this strategy creates a slight resistive loss, the overall efficiency remains robust at 87.41%. Therefore, $i_d^* = -25$ A is selected as the optimal point, effectively balancing the conflicting objectives of maximizing efficiency and ensuring superior power quality.

3.5. Comprehensive Performance Analysis

To validate the effectiveness of the selected strategy ($i_d^* = -25$ A), a direct comparison with the baseline ($i_d^* = 0$ A) is executed at the target high-speed condition of $\omega_m^* = 550$ rad/s.

The transient speed tracking performance, depicted in Figure 12, shows the superior dynamic response of the optimal strategy. The optimal approach achieves the target speed of 550 rad/s in 0.6 s, whereas the baseline method is slower, settling at 0.7 s. By injecting negative d -axis current, the effective rotor flux is reduced, enabling faster acceleration capabilities in the high-speed range.

The improved stability is explained by the voltage headroom analysis in Figure 13. Under the baseline condition, the voltage amplitude reaches 235.6 V, causing saturation. In contrast, the suggested FW control reduces the peak voltage to 207.8 V (a 27.8 V reduction). This creates a critical "voltage headroom" preventing the loss of current control.

Consequently, the impact of voltage stabilization on mechanical performance is clearly observed in Figure 14. While the proposed method presents a negligible increase in ripple at low speeds, its efficacy becomes decisive at high speed range. Specifically, as the speed reaches 600 rad/s, the baseline torque ripple spikes uncontrollably to 13.57 Nm. In contrast, the proposed strategy effectively suppresses this oscillation to just 4.97 Nm, yielding a net torque ripple reduction of 8.6

Nm, thereby supporting the high-speed operating capability by preserving control integrity and mechanical safety at the maximum tested speed.

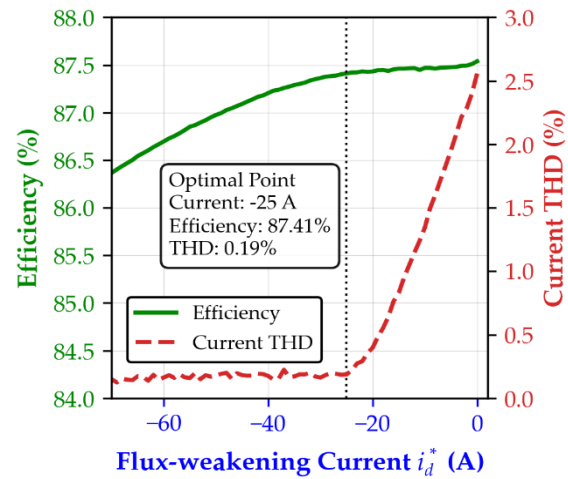


Figure 11. Effect of flux-weakening current on efficiency and THD performance

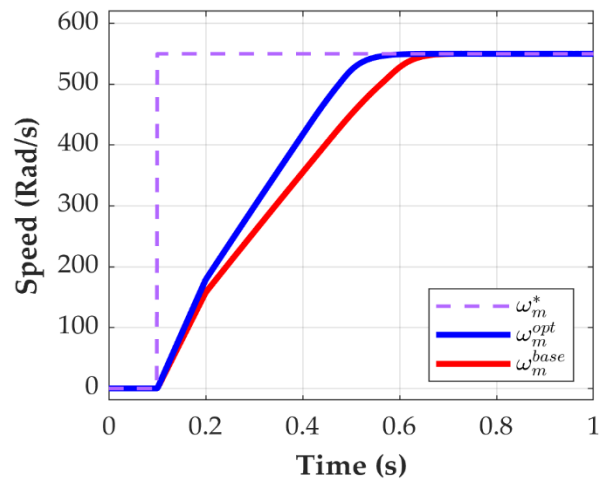


Figure 12. Speed tracking response comparison at 550 rad/s

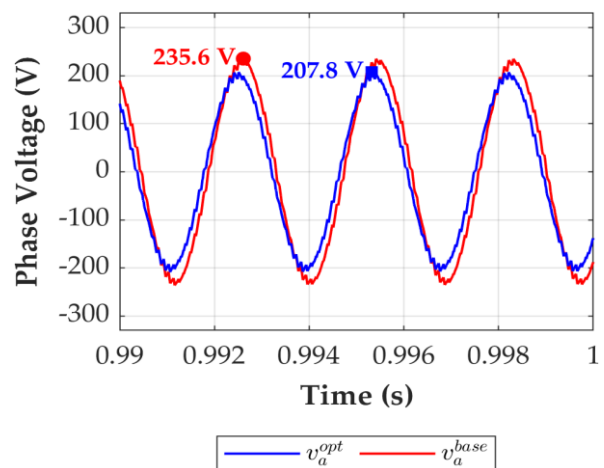


Figure 13. Phase voltage amplitude comparison at 550 rad/s

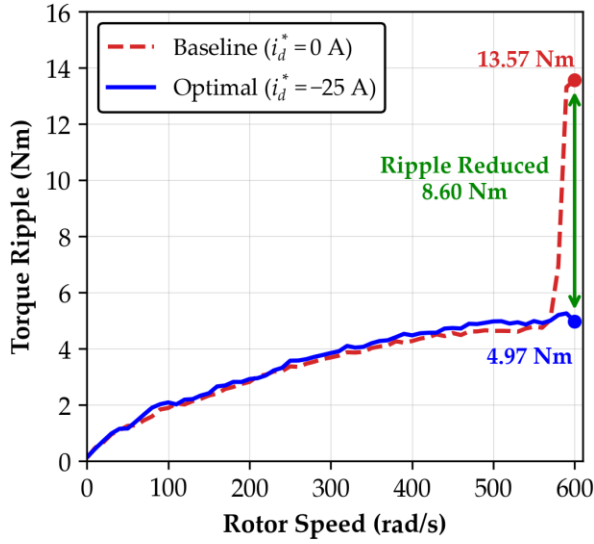


Figure 14. Torque ripple analysis across speed range

To elucidate the physical mechanism behind the ripple reduction, the harmonic spectrum of the stator currents is analyzed in Figure 15. The baseline control exhibits prominent low-order harmonics as a result of overmodulation. In particular, the 5th and 7th order harmonics are recorded at 2.426% and 0.743%, respectively. In contrast, the optimized strategy effectively suppresses these components, reducing the 5th harmonic to 0.066% and the 7th to 0.102%. Higher-order distortion is likewise attenuated; for example, the 11th order drops from 0.170% to 0.032%. This harmonic attenuation is directly consistent with the enhanced torque stability observed.

The stability is visually confirmed by the operating trajectory in the $d-q$ plane, shown in Figure 16. The baseline operating point is located at coordinates (0.0; 40.8), which falls outside the Voltage Limit Ellipse, indicating a loss of controllability. In contrast, the optimized operating point is shifted to (-25.0; 36.0). This coordinate lies safely within the safe operating area, which is the intersection of the Current Limit Circle and Voltage Limit Ellipse. This validates that the system operates strictly within its physical constraints.

3.6. Discussion

The analysis presented in Section 3.5 provides compelling evidence that the proposed optimized FW strategy ($i_d^* = -25$ A) significantly outperforms the baseline control across all critical performance metrics. These holistic

improvements are quantitatively summarized in Table 2 and visually synthesized in the multi-domain radar chart presented in Figure 17.

Regarding dynamic performance, the transient speed tracking improves by 14.3%, reducing the settling time from 0.7 s to 0.6 s. Unlike conventional schemes that suffer from saturation-induced instability and slower transient response [31], our proactive flux management maintains a higher dynamic bandwidth, ensuring faster response.

Crucially, stability is reinforced by an 11.8% reduction in peak phase voltage (from 235.6 V to 207.8 V). While existing studies often push the operation to the voltage limit to maximize power output [32], leading to potential loss of control, the created 27.8 V headroom serves as a vital safety buffer. This practically validates the theoretical stability boundaries proposed in [33], ensuring the current controller remains effective without hitting the inverter ceiling.

In terms of power quality, the method suppresses THD to a negligible 0.19%, a distinct advantage over typical overmodulation strategies which often exhibit 2–4% THD [34]. The 5th and 7th order harmonics are mitigated with reduction factors of 36 times and 7 times, respectively. This result is superior to the limit specified in the IEC 60034-1 standard (equivalent to TCVN 6627-1), which dictates that the Harmonic Voltage Factor (HVF) should not exceed 0.02 (2%) [8]. The system also ensures full compliance with the stringent IEEE 519 standard, which mandates a distortion limit of 5% for power systems [9].

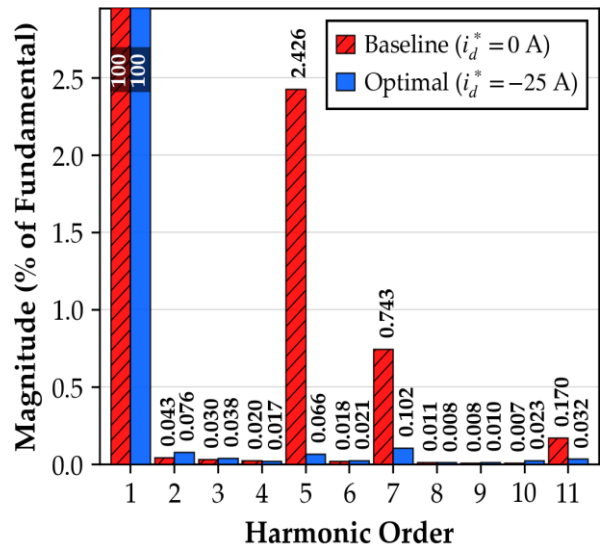
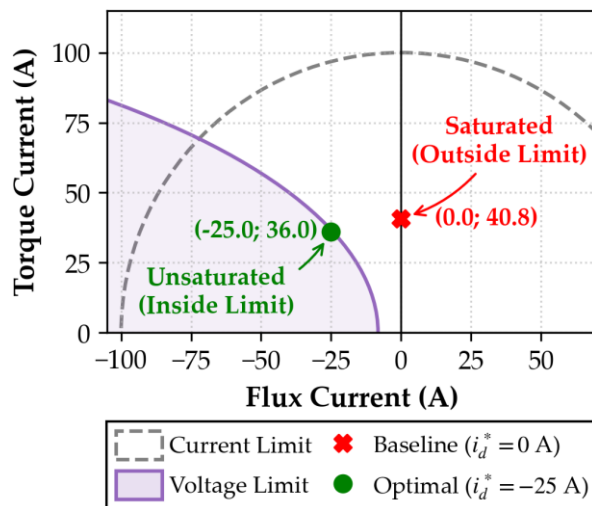
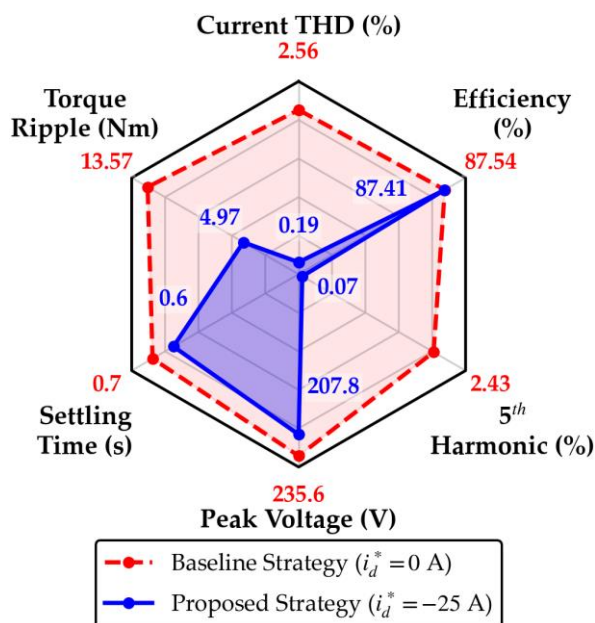


Figure 15. Harmonic spectrum comparison

Table 2. Performance comparison of control strategies at high-speed

Performance Metric	Baseline Strategy	Proposed Strategy	Improvement
Transient Settling Time	0.7 s	0.6 s	14.3% Faster
Peak Phase Voltage	235.6 V	207.8 V	27.8 V Headroom
Operating Coordinate	(0.0; 40.8)	(-25.0; 36.0)	Safe Zone
Total Harmonic Distortion	2.56%	0.19%	92.6% Reduction
5 th Harmonic Magnitude	2.43%	0.07%	Suppressed
Torque Ripple (600 rad/s)	13.57 Nm	4.97 Nm	8.6 Nm Reduction
System Efficiency	87.50%	87.41%	Negligible (< 0.1%)

**Figure 16.** Operation trajectory in d-q plane**Figure 17.** Comprehensive performance comparison of the proposed strategy against the baseline reference at high-speed operation

Finally, the coordinate shift from the unstable point (0.0; 40.8) to the optimized (-25.0; 36.0) provides definitive geometric proof of robustness. This trajectory correction aligns with recent research emphasizing that optimizing the current

vector path is essential to suppress harmonic excitations [35]. To explicitly address the correlation between these performance metrics, the improvements are evaluated against the baseline FOC without FW ($i_d^* = 0$ A) as the definitive reference. At the upper speed limit of 600 rad/s, this baseline strategy suffers from severe voltage saturation, which distorts the phase currents and amplifies the ripple to an unstable 13.57 Nm. By applying the optimized current ($i_d^* = -25$ A), the proposed strategy reclaims linear voltage modulation. This electrical enhancement yields a 92.6% reduction in THD to a negligible 0.19%, which directly translates mechanically by suppressing the torque oscillation by 8.6 Nm. Ultimately, it is this coupled suppression of electrical distortion and mechanical vibration guarantees control integrity, defining the true enhancement of the high-speed operating capability.

4. Conclusion

This study has established an optimized FOC strategy for IPMSM drives, effectively resolving the inherent conflict between operational speed extension and power quality in EVs powertrain. By utilizing parameters characteristic of a commercial C-segment EVs as a case study, the investigation confirmed that identifying a precise FW injection point ($i_d^* = -25$ A) provides a robust solution to the instability caused by rigid DC-bus voltage constraints. The research findings substantiate that creating an 11.8% voltage headroom is a prerequisite for maintaining controllability in the deep FW region. Establishing this safety buffer enhances the dynamic bandwidth by 14.3% while avoiding the instability associated with voltage saturation. Furthermore, geometrically correcting the operating trajectory into the safe area serves as a robust solution for mechanical vibration. This strategy reconciles electromagnetic performance with mechanical

durability by achieving a net torque ripple reduction of 8.6 Nm and ensuring full compliance with both IEC 60034-1 and IEEE 519 standards. These results contribute to the field by demonstrating that high-speed stability does not require sacrificing mechanical lifespan, offering a practical guideline for reducing vibration in modern EV powertrains. Future research will expand this architecture by integrating Reinforcement Learning algorithms to autonomously tune controller gains, thereby enhancing system adaptability to component aging and environmental fluctuations.

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Author's Declaration

Authors' contributions and responsibilities

Le Thanh Phuc, Le Ho Minh Khoa: The authors made substantial contributions to the conception and design of the study. Le Ho Minh Khoa: The author took responsibility for data analysis, interpretation, simulation, and discussion of results. Le Thanh Phuc: The author proposed the research methodology, modified and approved the manuscript.

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All data are available from the authors.

Competing interests

The authors declare no competing interest.

Generative AI statement

During the preparation of this manuscript, the author(s) used Google Gemini for English grammar and vocabulary correction. All AI-assisted output was subsequently reviewed and, where necessary, edited by the author(s), who takes full responsibility for the final content of the manuscript

Additional information

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